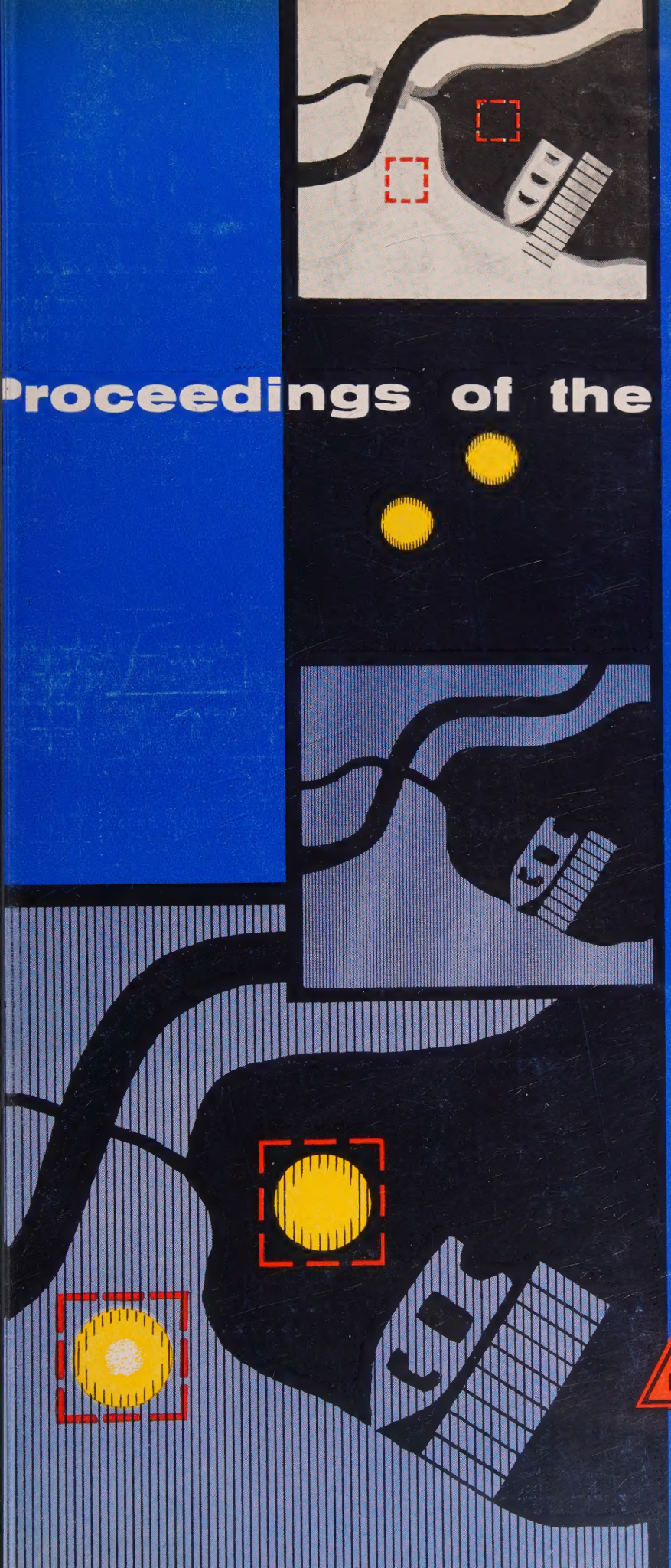

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Proceedings of the IRE

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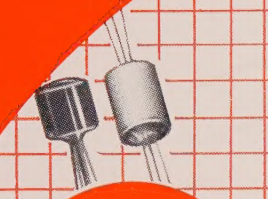
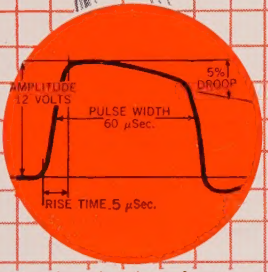
TWO-COLOR IMAGE INTENSIFIER: PAGE 1842

SPECIAL PULSE TRANSFORMERS TO YOUR REQUIREMENTS

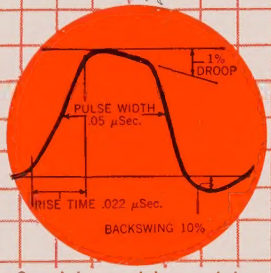
The pulse units illustrated below show a few of the thousands of special types designed and produced by UTC, to customers' requirements. Range covered on special pulse units is from a few microwatts to 10 megawatts. Rectangular pulse shapes are deliberately shown exaggerated to clarify parameters.

Almost thirty years of pioneering in the design and production of transformers plus exhaustive life testing programs and rigid quality control measures guarantee components of the highest reliability in the industry. . . . You can stake YOUR reputation on UTC products.

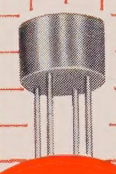


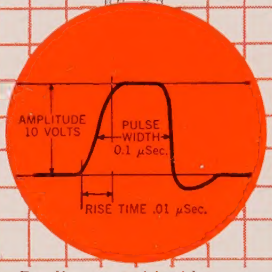


25 KC pulse transformer, DO-T or DI-T configuration. Pulse width 60 μSec. Rise time less than 0.5 μSec. Secondary C.T. balance each side to within 1% to ground. MIL-T-27A, GR 4. Size: DO-T, 5/16 dia. x 13/32", wt. 1/10 oz.; DI-T, 5/16 dia. x 1/4", wt. 1/20 oz.

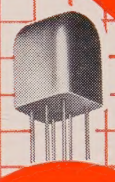


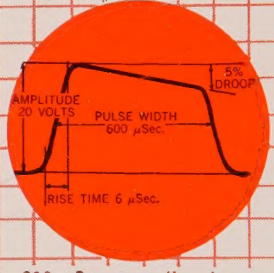
Special precision miniature pulse transformers. Designed in our standard stock mold to your specs. Checked and precisely adjusted in your tube or transistor blocking oscillator circuit. Sizes: 3/8 dia. x 3/8", 1 gram; 5/16 dia. x 7/16", 4 grams; 5/8 dia. x 5/8", 6 grams.



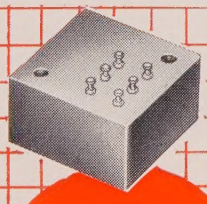


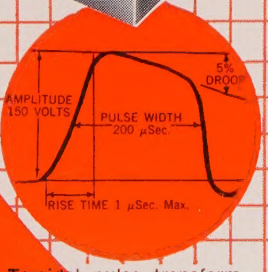
Ferrite core blocking oscillator transformer. 0.1 μSec. ±10% @ 200 KC PPS. 2 windings, rise time .01 μSec. Epoxy case. MIL-T-27A; 7/16 dia. x 1/4", .07 oz.



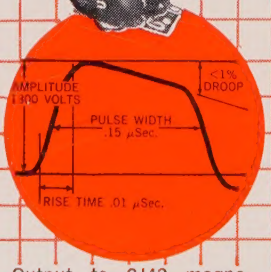


600 μSec. coupling transformer for printed circuit application. 3 windings, mu metal case for extreme shielding. Z=20 K Ω, 10 V. MIL-T-27A; standard UTC ML case; 7/16 x 31/64 x 9/16" h., .2 oz.

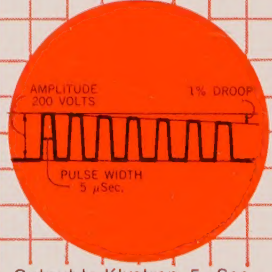




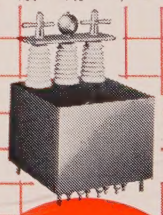
Toroidal pulse transformer, 150 V. 200 μSec. @ 400 PPS. Molded in epoxy. 3 windings, low leakage, less than 1 μSec. rise time; —40°C. to +85°C.; output voltage within 1%. MIL-T-27A; 1 3/8 sq. x 3/4"; 1.5 oz.

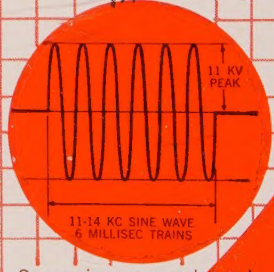


Output to 2J42 magnetron. Input 1300 V.-50 ohms. Output 6.5 KV to 1200 ohms and .6A. bifilar filament winding. .15 μSec., 1000 PPS. Trigger winding. MIL-T-27A GR 5; 1½ x 2½ x 2¾", 10 oz.



Output to Klystron, 5 μSec. pulses in groups of pulse trains at high rep rate. Droop 1% over pulse trains. 30 KV W.V., 43 KV hipot; —53°C. to +85°C.; MIL-T-27A; 4¼ x 5 x 6¾"; 11½ lbs.





Sonar sine wave pulse output transformer. PP 4-65A's, 11-14 KC flat. Pri. 11 KV; 28 KV hipot. Spark gap protected. Sec. 1500 V. @ 800Ω, 60 millisec on, 6% duty cycle. MIL-T-27A; —65°C. to +85°C.; 6 x 6 x 8½", 13 lbs.

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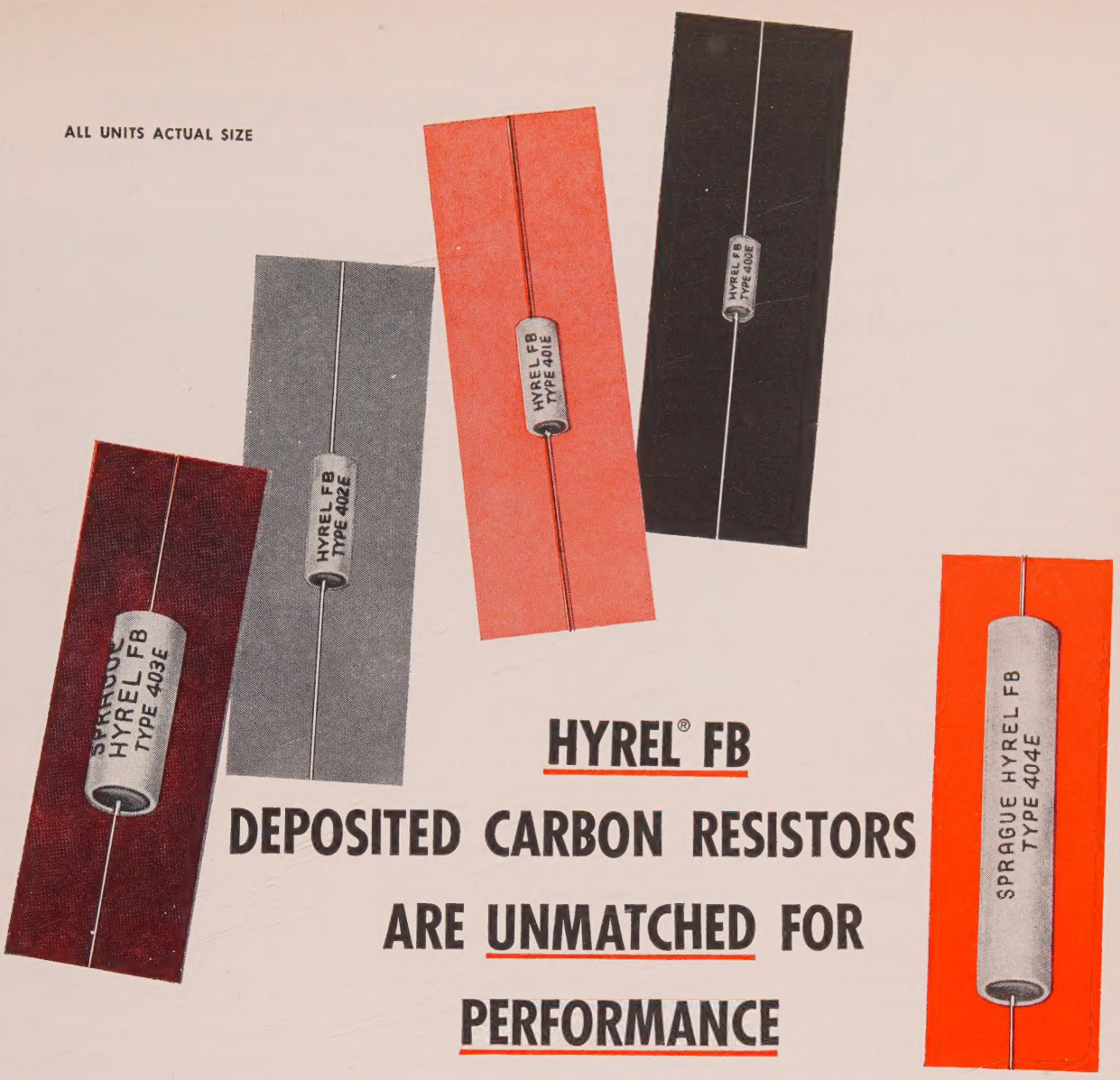

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In many applications there is need for a low-noise amplifier with a narrow amplification band width that can be tuned rapidly over a wide frequency range. Marty Grace and Sy Okwit of the Department of Applied Electronics have been working on a unique amplifier that may provide an exceptionally good solution to this problem. They present some of their preliminary results below.

Electronically Tunable Octave-Wide Parametric Amplifier with Constant Output Frequency*

In receiving systems used for frequency jumping radars, brief-signal detection, rapid analysis of low-level signal spectra, etc., it is extremely desirable to have a narrow-band receiver that (1) is capable of being electronically tuned across a wide frequency range, and (2) has a low noise figure. The conventional form of the backward-wave parametric amplifier (BWPA)[1,2] has these desired properties, but the center frequency of its output pass band, which is the same as that of the input pass band, varies as the amplifier is tuned. This requires a complex demodulator after the output port to convert the variable output frequency to a fixed intermediate frequency. Therefore, a more desirable form of this amplifier is one in which the center frequency of the output pass band remains constant as the input pass band is tuned. Such a mode of operation is obtainable, and since the tuning of the input pass band can be accomplished solely by electronically varying the pump frequency, the amplifier requirements stated above can be met quite simply.

In a typical BWPA, both the signal and idler frequencies are propagated on one transmission line, and the pump frequency propagates on a second transmission line that is coupled to the first through a distributed arrangement of nonlinear capacitance diodes. To obtain operation with an arbitrary circuit arrangement, the following general conditions must be met:

$$f_p = f_s + f_i \qquad (1)$$

$$\beta_p = \beta_s + \beta_i \qquad (2)$$

$$(V_g)_i \cdot (V_g)_s < 0 \qquad (3)$$

where f_p, f_s, and f_i are the respective frequencies of the pump, signal, and idler; β_p, β_s, and β_i (which can either be positive or negative) are the respective phase propagation constants at the pump, signal, and idler frequencies; and $(V_g)_s$ and $(V_g)_i$ are the respective group velocities at the signal and idler frequencies.

Equations 1 and 2 are identical with the conditions required for amplification in a traveling-wave parametric amplifier. Equation 3 determines whether the amplifier will operate in the traveling-wave or backward-wave mode. For the backward-wave mode, the group velocities of the signal and idler must be in opposite directions, making their product less than zero, whereas in the traveling-wave mode the group velocities must be in the same direction, making their product greater than zero.

* The work reported here was performed under Contract AF 30(602)-2197 with the Rome Air Development Center, Griffiss Air Force Base, N.Y.

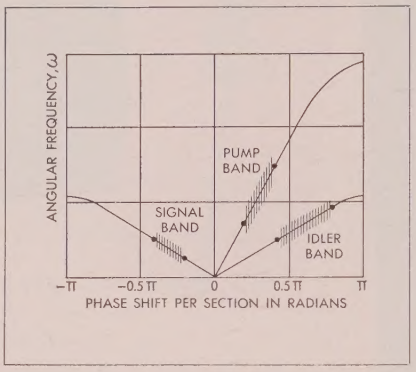

Figure 1. ω-β Diagram for Conventional BWPA

In a conventional BWPA, equations 1, 2, and 3 are satisfied by having the signal and idler waves propagate in opposite directions on a common low-pass line that is coupled to a second low-pass line on which the pump wave propagates in a direction opposite to that of the signal wave. These conditions are illustrated in the ω-β diagram of Figure 1.

In our case, the signal and idler waves are also propagated in opposite directions on a common low-pass line, but the pump wave propagates in a backward-wave mode on a band-pass line in the same direction as the signal wave. With this configuration, equations 1, 2, and 3 are also satisfied, but the idler frequency remains constant as the pump frequency is varied to tune the signal response of the amplifier. (See the ω-β diagram and block diagram of Figure 2.) Here, we can use the fixed idler as our output frequency, thereby eliminating the second-stage tuning and tracking problems.

This method of operation also provides the following unique advantages over conventional BWPA operation: (1) a high pump frequency can be used, resulting in a small percentage tuning range for the pump frequency even for octave-wide signal tuning, and (2) taking the useful output at the idler frequency gives the same theoretical noise performance as when it is taken at the signal frequency, but it produces a greater voltage gain-bandwidth product by a factor of about $\sqrt{f_i/f_s}$.

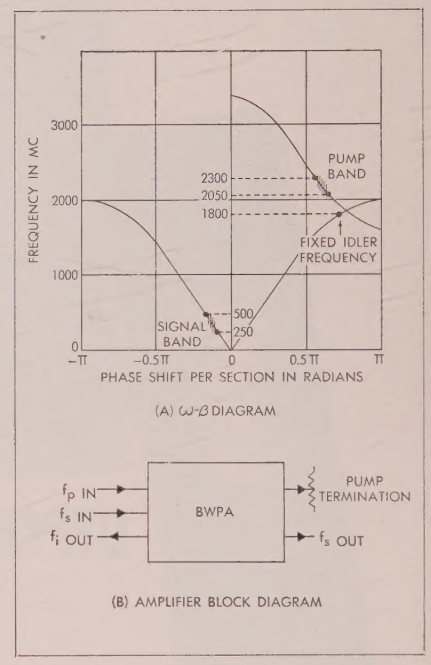

Figure 2. ω-β Diagram and Block Diagram for BWPA with a Constant Output Frequency

At present, we are working on such a device having the following design goals: (1) 250 to 500 Mc signal tuning in less than 100 microseconds (2) noise figure less than 2 db, and (3) voltage gain-fractional bandwidth product of at least 0.1. The pump frequency for this unit will vary from 2050 to 2300 Mc and the fixed idler frequency will be 1800 Mc.

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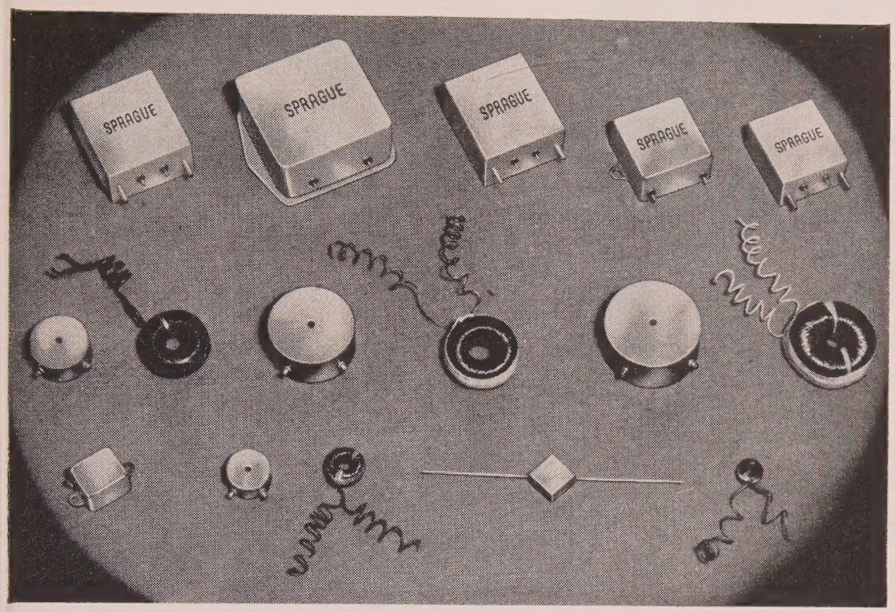


DESIGNED FOR USE in commercial, industrial, and military apparatus, Sprague Precision Toroidal Inductors are customarily supplied to the close inductance tolerance of ±1%. The broad line of Sprague Precision Toroidal Inductors includes such styles as open coil, plastic-dipped, rigid encapsulated inductors with tapped or through-hole mounting, and hermetically-sealed inductors.

All styles, with the exception of the open coil type construction, meet the appropriate requirements of Military Specification MIL-T-27A.

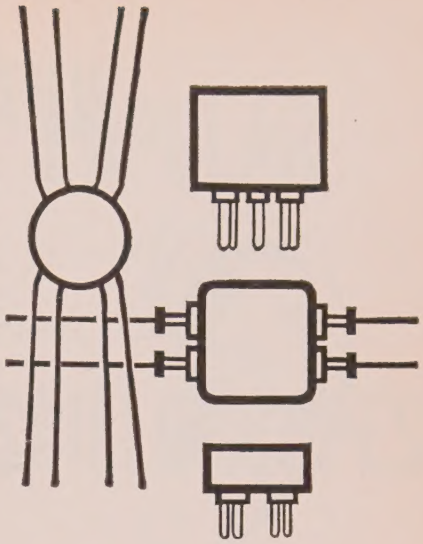
Sprague Precision Inductors are manufactured in modern plants which are equipped with the most up-to-date facilities for winding, processing, and testing the cores. Production instruments used in the manufacture of Sprague inductors are calibrated periodically to assure desired levels of accuracy. Quality control and inspection departments, which function independently of each other, maintain close surveillance over all production operations.

Several core permeabilities may be obtained in each of the five basic sizes of Sprague inductors to give the circuit designer the optimum selection of desired Q and current carrying abilities. Further, each of the core sizes is available with several degrees of stabilization. Inductors made with cores which have not been subjected to the stabilization process exhibit low inductance drift with time and have a low temperature coefficient of inductance. Where a greater degree of permanence of characteristics is required, cores with two different stabilization treatments can be used for most types of inductors.

All standard inductors by Sprague may be operated over the temperature range of –55 C to +125 C. Temperature cycling of finished inductors is a standard production procedure in order to equalize internal stresses and insure permanence of electrical characteristics.

In those cases where the extensive line of Sprague standard inductors is unsuitable for a particular application, the Special Products Division of the Sprague Electric Company will be glad to work with you to custom-tailor designs to meet specific customer requirements.

For detailed information on standard ratings, package sizes, Q, current carrying abilities, properties, etc., write on company letterhead for portfolio of engineering data sheets on precision toroidal inductors to Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.



Sprague offers standard and special

PULSE TRANSFORMERS

for military and commercial applications

Standard or special . . . military or commercial . . . in any size or shape . . . Sprague Pulse Transformers are designed to give top performance . . . in high-speed computer circuits, pulse inversion circuits, impedance matching circuits, blocking oscillator circuits, and many others.

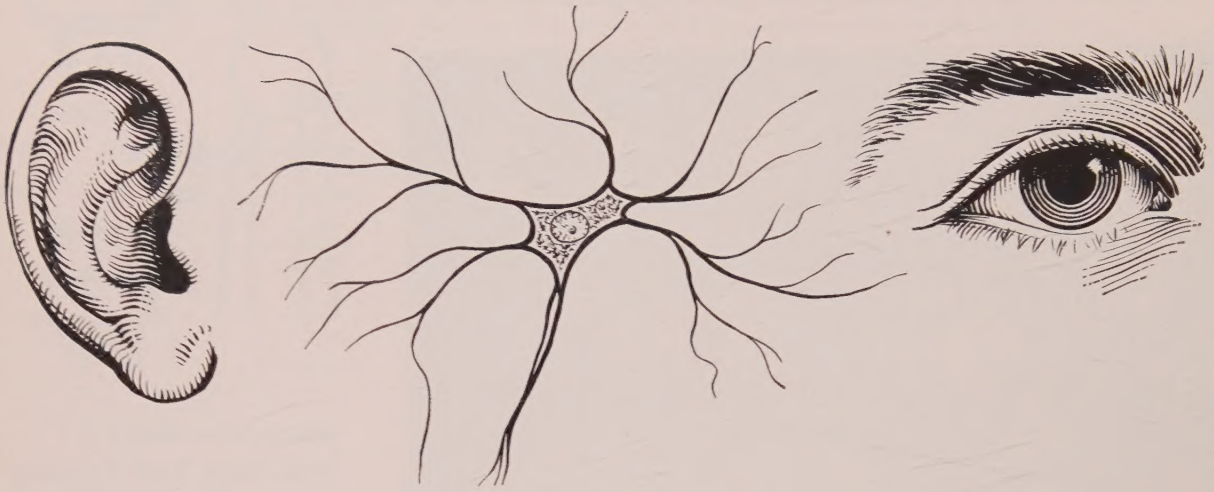
Special designs for high acceleration, high ambient temperatures or minified circuits can be furnished to suit requirements. Units are also available in lower-cost commercial housings.

For engineering assistance on your pulse transformer problems, write to Special Products Division, Sprague Electric Company, Union St., North Adams, Mass.

SPRAGUE®

THE MARK OF RELIABILITY

WHAT GOES ON HERE?



Bell Telephone Laboratories' new electronic "nerve cell" is a step toward finding out

One fascinating area of communications has long resisted exploration—what happens inside the nervous system when you see, or when you hear.

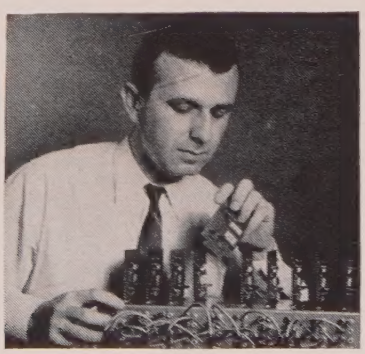
This area is of special interest to telephone science; knowledge of how the nervous system handles sound and picture signals can help determine what information is essential to perception. This in turn may lead to more efficient communication instruments and systems.

To probe the mystery of nerve activity, Bell Telephone Laboratories scientists have developed an electronic model of a living nerve cell or neuron. Consisting of transistors, resistors, capacitors and diodes, the "artificial neuron" exhibits many of the characteristics of a living neuron; for instance, "all-or-none" response and fatigue.

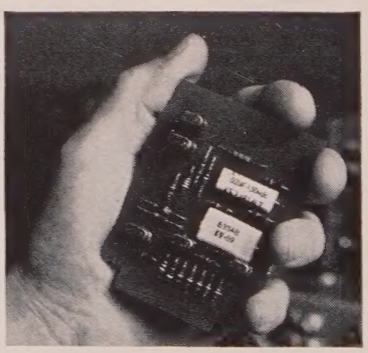
In one experiment at Bell Laboratories, a network of artificial neurons is subjected to a stimulus from light through a set of photocells. The network can distinguish specific patterns of light and dark, thus duplicating roughly some of the eye's basic reactions to light. Similar studies are underway to explore our hearing processes.

At present, too little is known about neural action to permit exact electronic duplication. But experiments with artificial neurons can provide suggestive clues, contributing to a stimulating interplay between electronics and neurophysiology which may help workers in both disciplines.

The human nervous system, including the brain, is the most efficient and versatile data processing system known; and data processing is an essential part of communications. The artificial neuron provides a new approach to investigating and understanding basic nerve network functions. It is a fresh example of how Bell Telephone Laboratories constantly explores new frontiers to improve America's communications system, now and in the years ahead.



Network of neurons is assembled by L. D. Harmon of Bell Laboratories, the initiator of this new research. Many kinds of assemblies are possible.



A single artificial neuron. It delivers electrical impulses when stimulated, like a living cell. Neurons are also being used for research into hearing.

BELL TELEPHONE LABORATORIES

WORLD CENTER OF COMMUNICATIONS RESEARCH AND DEVELOPMENT

Call your hp rep today for a demonstration of one of these

3 POPULAR hp OSCILLOSCOPES

Production or lab instruments—Simple to use, even for non-technical personnel—Moderately priced—Full 10 cm x 10 cm display—Automatic calibration waveforms—Low phase shift—Automatic triggering for optimum presentation—"Times-5" sweep expander magnifies trace, improves resolution.

DC to 200 KC

Models 120A/AR combine minimum controls with hp automatic triggering for utmost speed, convenience. Horizontal amplifier dc to 200 KC; phase shift only ± 2° to 100 KC. More X-axis information due to horizontal amplifier sensitivity control, with vernier, 5% accuracy. Balanced input on most sensitive ranges for low level work. Times-5 sweep expander, all ranges. 15 calibrated sweep speeds, 5 μsec/cm to 0.2 sec/cm. Vernier, expander extend speed range 1 μsec/cm to 0.5 sec/cm. 10 mv/cm sensitivity calibrated vertical amplifier, drift-free trace. hp 120A (cabinet) or hp 120AR (rack), $435.

DC to 200 KC–DUAL TRACE

Models 122A/AR provide simultaneous two-phenomena presentation, are ideal for direct comparison of filter, amplifier output/input phenomena; vibration testing. Unique hp front-panel automatic calibrator waveform switch. Twin vertical amplifiers operate independently, simultaneously, differentially. Automatic triggering, automatic synchronization, single trace operation when desired. Sensitivity 10 mv/cm to 100 v/cm, 15 calibrated sweeps, vernier extension. Horizontal amplifier dc to 200 KC. hp 122A (cabinet) or 122AR (rack), $625.

DC to 300 KC–"BIG SCOPE" PERFORMANCE

Models 130B/BR provide wide usefulness, simple operation and rugged dependability. 21 calibrated sweep times, 1 μsec/cm to 5 sec/cm. Vernier, expander extend range 0.2 μsec/cm to 12.5 sec/cm. Twin horizontal and vertical amplifiers, phase shift ± 1° to 50 KC; sensitivity 1 mv/cm to 125 v/cm. Balanced input on 6 most sensitive ranges. Common mode rejection 40 db. Stability 1 mv/hour after warmup. Triggering automatic, internally, line power, externally, 0.5 v or greater. hp 130B (cabinet) or 130BR (rack), $650.

Data subject to change without notice. Prices f.o.b. factory.

Thirteen precision hp oscilloscopes, dc to 1,000 MC

HEWLETT-PACKARD COMPANY

1050D Page Mill Road, Palo Alto, California, U.S.A.
Cable "HEWPACK", DAvenport 6-7000

Sales representatives in all principal areas

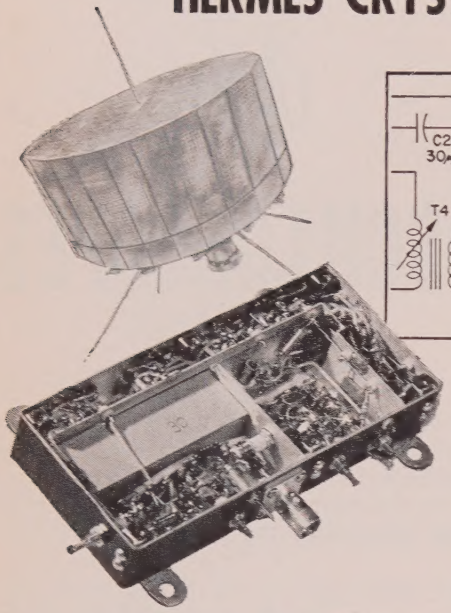
HEWLETT-PACKARD S.A.
Rue du Vieux Billard No. 1, Geneva, Switzerland
Cable "HEWPACKSA", Tel. No. (022) 26. 43. 36

6715

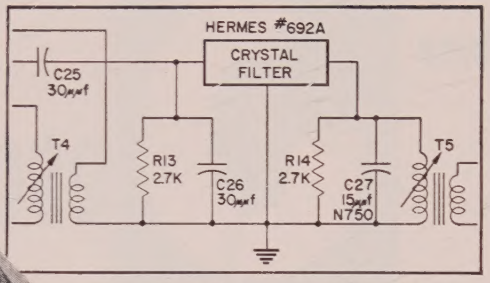
FIRST Tiros Weather Satellite

Uses

HERMES CRYSTAL FILTERS



One of the two Command Receivers in TIROS



A portion of the schematic diagram for the Command Receiver in TIROS showing Hermes 20 Mc Crystal Filter at first mixer.

Hermes Crystal Filter, Model 692A, shown half size, occupies only 1.5 cu. ins.

The TIROS satellite, carrying the nation's most advanced space-borne television "eye" to study the world's weather, comprises perhaps the most elaborate electronics package yet sent into orbit.

The information-gathering element in a complex satellite-and-ground system developed for the National Aeronautics and Space Administration by RCA, TIROS contains miniature TV cameras, video tape recorders, transmitters, solar cell and re-chargeable battery power supplies, and an array of control and communications equipment. One Hermes Crystal Filter, Model 692A, is used in each of two Command Receivers which pick up coded signals transmitted from the ground to establish the time during orbit when cameras, tape recorders, and playback equipment will operate. Launched April 1, 1960, TIROS relayed meteorological data for 78 days.

Hermes Crystal Filters were selected because of their unusual ability to meet the severe environmental conditions encountered in space, while providing extremely high selectivity to receive command signals reliably. Characteristics of Hermes Crystal Filter, Model 692A, include: *Center Frequency:* 20 mc ± 1 kc; *6db Bandwidth:* 40 kc min; *Passband Response Variation:* ±½ db; *60 db Bandwidth:* 100 kc max; *Impedance:* 1K nominal; *Midband Insertion Loss:* 3 db max; *Size:* 1.5 cu. ins; *Environment: Shock:* 100 G's; *Vibration:* 20 G's −2000 cps; *Center Frequency Variation:* ±2 kc over the *Temperature Range:* −55° C. to + 85° C.

If you have a filtering problem, call on Hermes engineering specialists to assist you in the design of your circuitry and in the selection of filter characteristics best suited to your needs. *Write for Crystal Filter Bulletin to Hermes Electronics Co., Dept B, 75 Cambridge Parkway, Cambridge 42, Mass.*

A limited number of opportunities are available to experienced circuit designers. Send Résumé to Dr. D. I. Kosowsky.

Hermes

ELECTRONICS CO.

75 CAMBRIDGE PARKWAY, CAMBRIDGE 42, MASS.

A DIVISION OF Itek

Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.

Δ

November 14-16, 1960

Mid-America Electronics Convention (MAECON), Hotel Muehlebach, Kansas City, Mo.

Exhibits: Mr. Gustav Vasen, H. A. Roes Co., 2106 Cherry, Kansas City, Mo.

November 15-17, 1960

Northeast Electronics Research & Engineering Meeting (NEREM), Boston Commonwealth Armory, Boston, Mass.

Exhibits: NEREM, 313 Washington St., Newton, Mass.

December 1-2, 1960

PGVC Annual Meeting, Sheraton Hotel, Philadelphia, Pa.

Exhibits: Mr. E. B. Dunn, Atlantic Refining Co., 260 S. Broad St., Philadelphia 1, Pa.

December 13-15, 1960

Eastern Joint Computer Conference, Hotel New Yorker & Manhattan Center, New York, N.Y.

Exhibits: Mr. Alan D. Meacham, 120 E. 41st St., New York 7, N.Y.

February 1-3, 1961

Second Winter Convention on Military Electronics, Biltmore Hotel, Los Angeles, Calif.

Exhibits: Mr. John Francis O'Halloran, O'Halloran & Associates, 11636 Ventura Blvd., North Hollywood, Calif.

March 20-23, 1961

International Radio and Electronics Show and IRE International Convention, Waldorf-Astoria Hotel and New York Coliseum, New York, N.Y.

Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 72 West 45th Street, New York 36, N.Y.

April 19-21, 1961

SWIRECO, South West IRE Regional Conference & Electronics Show, Dallas, Tex.

Exhibits: Mr. R. W. Olson, Texas Instruments, Inc., 6000 Lemmon Ave., Dallas 9, Tex.

April 26-28, 1961

Seventh Region Technical Conference and Trade Show, Westward Ho Hotel, Phoenix, Ariz.

Exhibits: Mr. G. J. Royden, 912 W. Linger Lane, Phoenix, Ariz.

May 8-10, 1961

National Aeronautical Electronics Conference (NAECON), Miami & Biltmore Hotels, Dayton, Ohio

Exhibits: Mr. Edwin P. Turner, Wright Patterson Air Force Base, Dayton, Ohio

(Continued on page 10A)

KAY *Ligna-Sweep*® *SKV*

SWEEPING OSCILLATOR

Cat No. 935-A

AUDIO—VIDEO—VHF . . . IN ONE INSTRUMENT!

- FREQUENCY RANGE—200 CPS TO 220 MC.
- SWEEP REPETITION RATES FROM 0.2 TO 60 CPS.
- LINEAR AND LOGARITHMIC SWEEPS.
- AUDIO SWEEP—200 CPS TO 20,000 CPS.
- 3 Highly Stable Video Bands—1 kc to 12 mc., Variable or in Single Sweep.
- RF Output of 1 Volt rms into 70 Ohms ±.5 db Over Widest Sweep Width.
- 8 Narrow Customer Selected Fixed Frequency Bands—20 kc to 12 mc.
- 9 Fundamental Frequency, Wide VHF Bands—10 mc to 220 mc.

SPECIFICATIONS

VARIABLE FREQUENCY RANGES: .5-12 mc, .1-12 mc, 10 kc-12 mc, 10-220 mc (9 bands)

FIXED FREQUENCIES: Up to max. of 8 center frequencies (20 kc to 12 mc—Customer selected)

AUDIO RANGE: 200 cps to 20 kc.

SWEEP WIDTHS: Selected for maximum stability 1-10 mc on .5-12 mc band; .2-2 mc on .1-12 mc band; 20-200 kc on 10 kc-12 mc band; 6% to 60% of center freq. to 50 mc and 3 mc to 30 mc above 50 mc on 10-220 mc bands. 2-20 kc on fixed frequencies and audio range.

OUTPUT LEVEL: Continuously variable from 1 volt rms down to 65 db below 1 volt, ±5% over widest sweep. AGC. Audio range: variable .5-1 volt rms.

IMPEDANCE: 70 ohms nominal (50 ohms on request). Audio range: 600 ohms.

SWEEP OUTPUT and REPETITION RATES: Sawtooth for horizontal deflection of oscilloscope. Approx. 7 volts peak to peak—Output Impedance 1000 ohms nom.; fixed 60 cps, line locked; fixed 30 cps. logarithmic (for audio and video application) 3 cont. var. ranges—.2-1 cps, 1-5 cps, 5-30 cps.

MARKERS: Swept signal available for operation of *Vari-Marker SKV* Generator.

Optional Internal Markers. Limited number of sharp, crystal-controlled pulse-type markers at customer specified frequencies can be provided. Please inquire before ordering.

POWER SUPPLY: Input approx. 220 Watts, 117v (±10%), 50-60 cps. B+ electronic regulation.

PRICE: $1295.00 f.o.b. factory. Fixed freq. bands add $17.00 per band.

The wide range of frequency and repetition rate in the *Ligna-Sweep Model SKV* make it ideally suited for alignment and testing of a wide variety of electronic instruments—audio amplifiers, filters, communication receivers, radar IF channels, TV receivers and transmitters.

The unit is stable and carefully shielded and filtered to prevent spurious signals on beat frequency video bands. A wide range of sweep repetition rates makes viewing easy on conventional oscilloscopes. Low repetition rates used with long persistence screens permit study of high Q circuitry, LF limits of band circuits and observation of the "ring" characteristics of tuned circuits.

Ligna-Sweep® MODEL CP

SWEEPING OSCILLATORS

Cat. No. 932-A: Variable bands between 100 kc and 215 mc. Price: $795.00 f.o.b. factory, 18 pulse-type markers avail. at customer specified freq., $17.00 each.

Cat. No. 932-B: Continuous coverage from 100 kc—150 mc. Price: $795.00 f.o.b. factory, 18 pulse-type markers avail. at customer specified freq., $17.00 each.

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KAY ELECTRIC COMPANY

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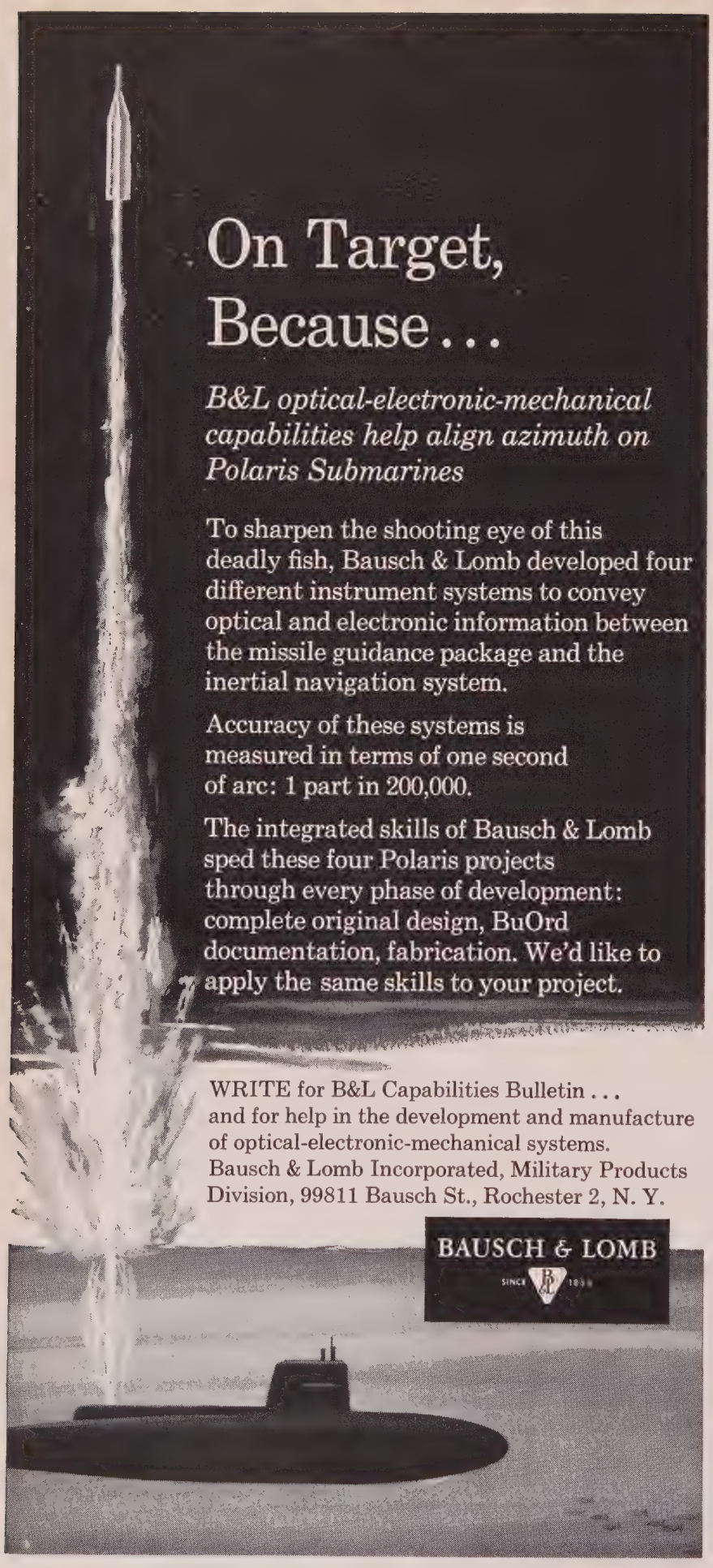
On Target, Because . . .
B&L optical-electronic-mechanical capabilities help align azimuth on Polaris Submarines
To sharpen the shooting eye of this deadly fish, Bausch & Lomb developed four different instrument systems to convey optical and electronic information between the missile guidance package and the inertial navigation system.
Accuracy of these systems is measured in terms of one second of arc: 1 part in 200,000.
The integrated skills of Bausch & Lomb sped these four Polaris projects through every phase of development: complete original design, BuOrd documentation, fabrication. We'd like to apply the same skills to your project.
WRITE for B&L Capabilities Bulletin . . . and for help in the development and manufacture of optical-electronic-mechanical systems. Bausch & Lomb Incorporated, Military Products Division, 99811 Bausch St., Rochester 2, N. Y.
BAUSCH & LOMB
SINCE 1853

Meetings with Exhibits

(Continued from page 8A)

May 9-11, 1961

Western Joint Computer Conference, Ambassador Hotel, Los Angeles, Calif.

Exhibits: John H. Whitlock Associates, 253 Waples Mills Road, Oakton, Va.

May 22-24, 1961

Fifth National Symposium on Global Communications (GLOBECOM V), Sherman Hotel, Chicago, Ill.

Exhibits: Mr. Robert Hajek, Commtronix Engineering, Box 62, Riverside, Ill.

June 6-8, 1961

Armed Forces Communications & Electronics Show, Sheraton Park and Shoreham Hotels, Washington, D.C.

Exhibits: Mr. William C. Copp, 72 W. 45th St., New York 36, N.Y.

June 26-28, 1961

Fifth National Convention on Military Electronics, Shoreham Hotel, Washington, D.C.

Exhibits: Mr. L. David Whitelock, BuShips, Electronics Div., Dept. of Navy, Washington, D.C.

July 16-22, 1961

Fourth International Conference on Bio-Medical Electronics & Fourteenth Conference on Electronic Techniques in Medicine & Biology, Waldorf-Astoria Hotel, New York, N.Y.

Exhibits: Mr. Lewis Winner, 152 W. 42nd St., New York 36, N.Y.

August 22-25, 1961

Western Electronic Show and Convention (WESCON), Cow Palace and Fairmont Hotel, San Francisco, Calif.

Exhibits: Mr. Don Larson, WESCON, 1435 LaCienega Blvd., Los Angeles, Calif.

October 4-6, 1961

IRE Canadian Convention, Exhibition Park, Toronto, Canada

Exhibits: Business Manager, IRE Canadian Convention, 1819 Yonge St., Toronto 7, Ontario, Canada

October 9-11, 1961

National Electronics Conference, Hotel Sherman, Chicago, Ill.

Exhibits: Mr. Arthur H. Streich, National Electronics Conference, 228 N. LaSalle St., Chicago 1, Ill.

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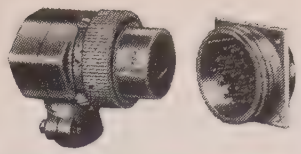
Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.

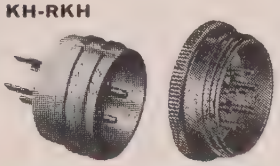
RELIABLE MULTI-PURPOSE "K" SERIES

K STANDARD



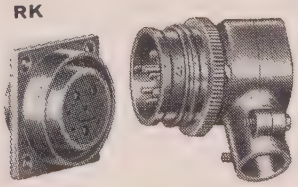
FOR AIRCRAFT, ELECTRONIC, INSTRUMENT, MILITARY, MISSILE, INDUSTRIAL AND COMMERCIAL APPLICATIONS
Standard K and RK . in straight and angle 90° plugs with wall mounting receptacles. Conduit and clamp entry types. 1 to 110 contacts in 250 different insert arrangements. 10, 15, 30, 40, 60, 80, 115 and 200 amp. silver plated brass on copper contacts. High quality phenolic, melamine, and formica insulators. Cadmium-plated aluminum alloy shells. Flashover voltages: 110 to 5000V, 60 cps ac rms.

KH-RKH



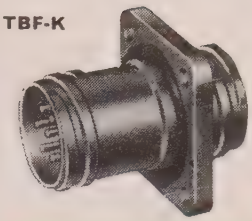
FOR USE UNDER CRITICAL PRESSURE AND LEAKAGE CONDITIONS—Hermetically sealed plugs with steel shells, steel contacts, and Canseal glass insulators for true hermetic sealing. Electro tin plating over cadmium plate over copper flash provides highly receptive surface for soldering and corrosion resistance.

RK



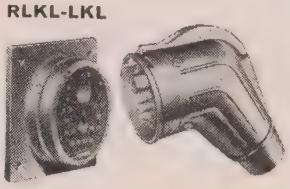
FOR FLUSH OR SEMI-FLUSHMOUNTING—Cannon RK Plug assemblies are equipped with an external threaded coupling nut which is the reverse of the standard K Series. Note: RK will mate only with RK's.

TBF-K



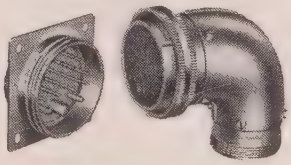
FOR CARRYING CIRCUITS THROUGH BULKHEADS
Cannon TBF-K Bulkhead Plugs feature a double-faced construction allowing mating at both ends. Pin inserts. Single piece shell.

RLKL-LKL



FOR TV AND OTHER PANEL SWITCHING OPERATIONS—Quick connect and disconnect RLKL Plugs are designed for one-hand fast disconnect use on TV station program switching panels and similar type operations. Feature a quick coupling means. Latch-lock secures plug to mated fitting (RLKL receptacle). Thumb pressure releases it.

FW-K FWR-K



FOR OPEN FLAME PROTECTION AGAINST HIGH TEMPERATURES—Cannon K Firewall Plugs are available in straight and angle 90° shell types. Wall mounting receptacles also available. Phenolic or fireproof inserts of glass-filled materials. Crimp type contacts. Cannon originated the firewall connector and continues to be the leader in this important field.

SK-M7-21C

FOR USE IN TELEPHONE "BEEPER" AND SIMILAR APPLICATIONS—Widely used on telephone recording units known as "beepers." Adaptable for other similar applications.

K ACCESSORIES—Cannon K Series Accessories include Straight and Angle 90° Junction Shells, Dust Caps, Bonding Rings, Gland Nuts, Clamps, and Dummy Receptacles to hold and protect plugs when not in use.

Cannon has available a wide variety of other Plug designs FOR EVERY CONCEIVABLE APPLICATION...including aircraft and electronic Plugs conforming to Specification MIL-C-5015D; Unit-Plug-In Rack/Panel and Modular; Audio and Low-Level Circuit Plugs; Miniature and Sub-Miniatures; Coaxial RF Series Plugs, Printed Circuit Plugs; GM Plugs and Cannon Plug/Harness Systems: "Kwik-Term" Terminals and DC Solenoids.

FOR ADDITIONAL INFORMATION on the typical designs illustrated...other configurations for your specific applications...or the design, engineering and manufacture to your special needs...write to **Cannon Electric Company**—3208 Humboldt Street, Los Angeles 31, California. Please refer to Dept. 841.

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CANNON PLUGS

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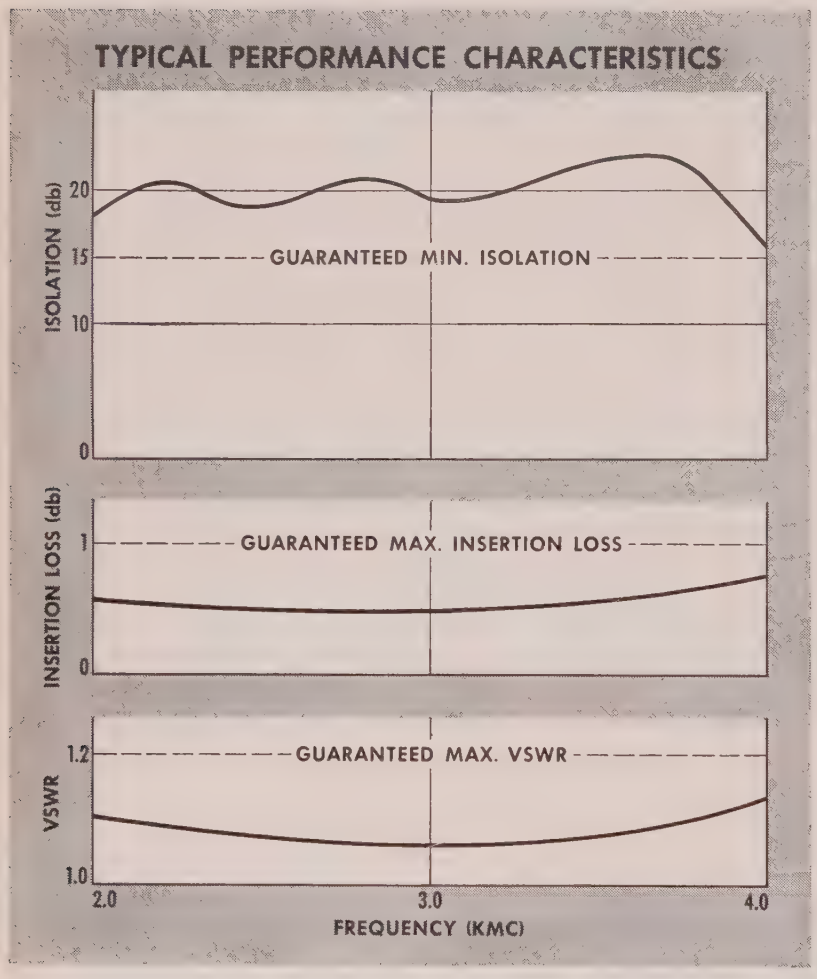
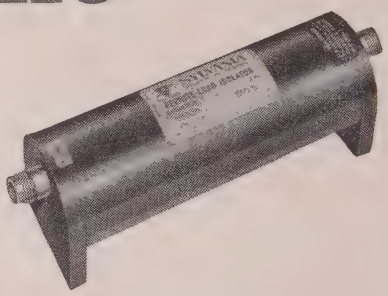
Microwave Component News from SYLVANIA

Octave Bandwidth Coaxial Isolators from 1-11 kmc





BROADBAND coaxial ferrite load isolators from 1-11 kmc (with octave steps from 1 to 8) are now available from Sylvania for almost every microwave application. They are especially well suited to test equipment and other wide band applications.

Operating curves for an FD-151P, shown at left, illustrate the outstanding performance of these units. In each case, operation exceeds the rated limits by a substantial margin throughout the entire band—not just at mid-band.

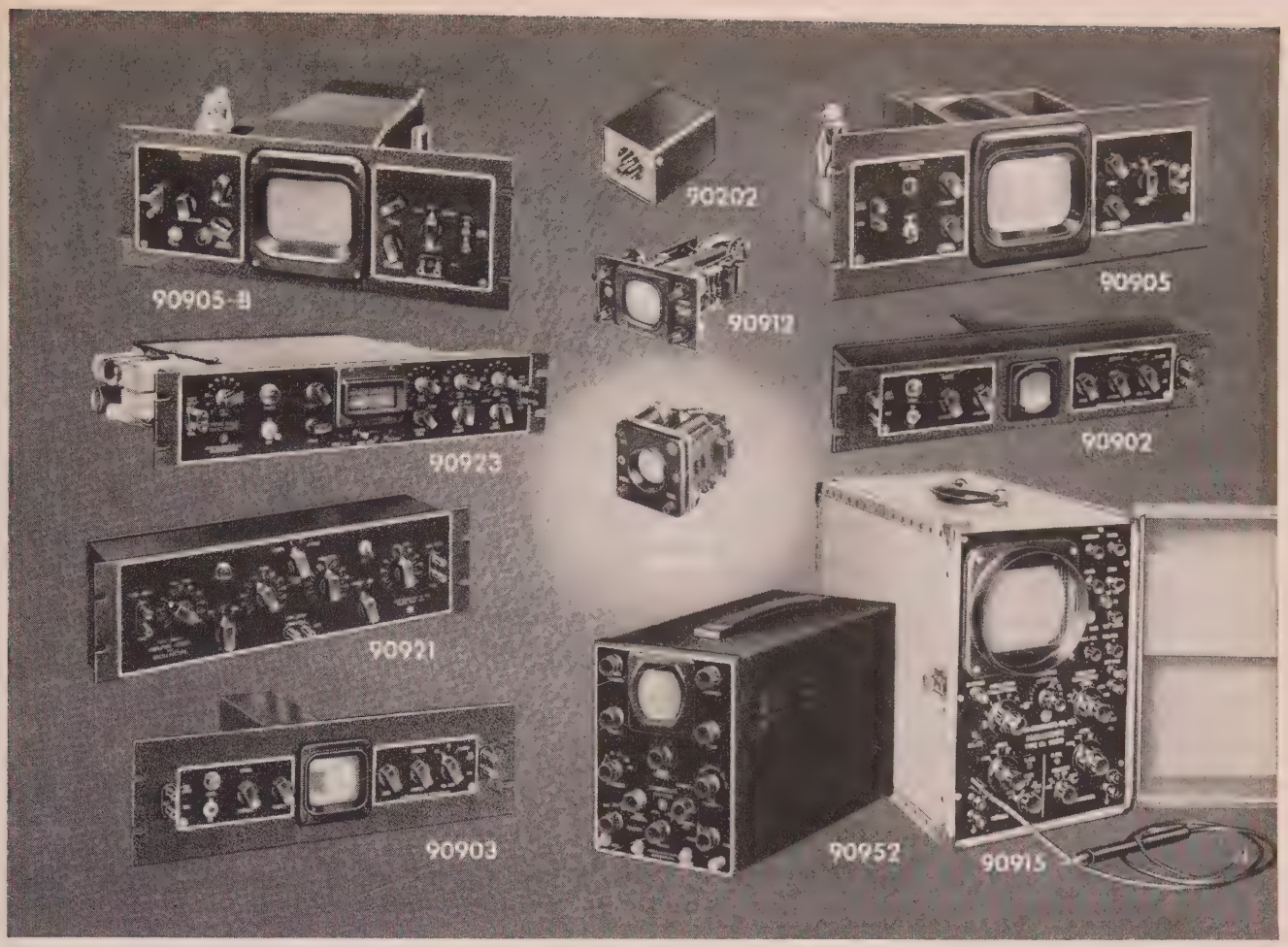
With normal handling, Sylvania ferrite isolators give this same electrical performance for years without deterioration or failure. They can be used to reduce the VSWR presented by a load or antenna, and to protect oscillator output from long line effects.

For more information on these units or other standard or custom built devices in Sylvania's extensive ferrite device line write, wire or phone your nearest Sylvania tube sales office, or contact Sylvania Special Tube Operations, 500 Evelyn Avenue, Mountain View, California.

SYLVANIA

Subsidiary of GENERAL TELEPHONE & ELECTRONICS

GENERAL SYSTEM



"Designed for Application" • Oscilloscopes and Accessories

The extensive Millen "Designed for Application" line of oscilloscopes and accessories includes five instrumentation oscilloscopes, six rack mounted basic oscilloscopes, an insulated industrial oscilloscope, a miniature synchroscope-oscilloscope, two compact rack mounted complete oscilloscopes, a rack mounted basic oscilloscope for military applications, two amplifier-sweeps, and a plug-in power supply.

INSTRUMENTATION OSCILLOSCOPES

Miniaturized, packaged panel mounting cathode ray oscilloscopes designed for use in instrumentation in place of conventional "pointer type" moving coil meter. Magnitude, phase displacement, wave shape, etc. are readily displayed.

No. 90901 uses type 1CP1 fixed focus one inch cathode ray tube. 2 11/32" x 2¾" panel. Panel bezel matches in size and type the standard 2" square meters.

No. 90911 uses Type 1EP1 cathode ray tube. Balanced deflection. Blanking input. Sharp focus. Panel matches 2" square meters. Flat face RCA 1¼" diameter tube.

No. 90912 uses type 2BP1 two inch cathode ray tube. 3" x 5" panel. Sharp focus. Good sensitivity. Accelerating voltage 500 to 875 volts. Min. control interaction.

No. 90912-R uses type 3UP1 2⅜" x 1⅝" rectangular cathode ray tube.

No. 90913 uses type 3XP1 3" x 1½" rectangular cathode ray tube. 1⅛" x 2¾" useful scan. Vertical sensitivity 33 volts d.c. per inch at 2000 v. accelerating.

BASIC OSCILLOSCOPES

Rack mounted inexpensive *basic* oscilloscopes including cathode ray tube circuit, power supply; intensity, focus, and centering controls, magnetic shielding, safety features, switches, etc. The *basic* oscilloscopes in their packaged form are entirely adequate for many laboratory as well as industrial and communication uses.

No. 90902 uses type 2BP1 two inch cathode ray tube. 3½" x 19" panel. Power supply — 105–125 volts — 60 cycles. Power consumption — 19 watts.

No. 90902-M — Military version of 90902.

No. 90903 uses type 3KP1 three inch cathode ray tube. 5¼" x 19" panel. Power supply — 105–125 volts — 60 cycles. Power consumption — 19 watts.

No. 90903-R uses type 3XP1 3" x 1½" rectangular cathode ray tube. 3½" x 19" panel. Power supply 105–125 volts — 60 cycles.

No. 90905 uses type 5UP1 five inch cathode ray tube. 7" x 19" panel. Power supply 105–125 volts — 60 cycles. Power consumption — 32 watts.

No. 90905-B uses type 5ADP1 five inch flat face precision tolerance cathode ray tube. Power supply — 105–125 volts — 60 cycles. Power consumption — 35 watts.

No. 90905-R uses type B1204 4⅝" x 2⅝" rectangular cathode ray tube. 3½" x 19" panel.

INDUSTRIAL OSCILLOSCOPE

Suitable for use in factory, laboratory, and the field for design, installation, maintenance, and service. Completely insulated front panel and case. Double shielded against magnetic fields. Excellent linearity. Sharp focus over entire 4" x 4" useful scan. The vertical and horizontal amplifiers are stable d.c. amplifiers and are identical, thus permitting accurate phase measurements.

No. 90915 uses type 5AQP — 1, 2, 7, or 11 flat face, precision tolerance cathode ray tube. Frequency response of either amplifier D.C. to 100 K.C. + 0–10%.

MINIATURE SYNCHROSCOPE-OSCILLOSCOPE

Compact "field service" Synchroscope or Oscilloscope. 7½" x 5⅝" x 13". Weighs 17 pounds. Synchronizes to internal or external positive or negative pulses. Band width 10 cycles to 1000 KC. Sweep 6 to 300 microseconds per inch. Performance has not been sacrificed in designing this unit for light weight.

RACK MOUNTED OSCILLOSCOPES

Complete with amplifiers and sweep. Good low frequency response and linearity. For monitoring, production test, or laboratory use. Compact.

No. 90923 uses type 3XP1 3" x 1½" rectangular cathode ray tube. 3½" x 19" panel.

No. 90925 uses type B1204 4⅝" x 2⅝" rectangular cathode ray tube. 3½" x 19" panel.

PLUG-IN POWER SUPPLY

Compact high voltage power supply for oscilloscopes, etc. 2" x 2½" x 5". Input 117 volts 50/60 cycles at 10 watts. Output 750 volts d.c. at 3 ma. and 6.3 volts a.c. at 600 ma. Supplies accelerating and centering potential for oscilloscopes.

No. 90202 Power Supply for Instrumentation Oscilloscopes.

AMPLIFIER/SWEEP UNITS

Horizontal and vertical amplifiers and sawtooth sweep generator for use with *basic* oscilloscopes. Match MILLEN *basic* oscilloscopes in appearance.

No. 90921 — 6SJ7 amplifiers. 6SN7-GT hard tube sweep. 5¼" x 19" rack panel.

No. 90922 — 3½" x 19" rack panel. Good low frequency linearity.

JAMES MILLEN MFG. CO., INC.

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IRE News and Radio Notes

Current IRE Statistics

(As of September 30, 1960)
Membership—82,559
Sections*—107
Subsections*—28
Professional Groups*—28
Professional Group Chapters—273
Student Branches†—192

* See this issue for a list.
† See October, 1960 for a list.

Calendar of Coming Events and Authors' Deadlines*

1960

6th Ann. Conf. on Magnetism and Magnetic Materials, New Yorker Hotel, N. Y., N. Y., Nov. 14–17. (DL*: Aug. 26. A. M. Clogston, R. C. Fletcher, Bell Tel. Labs., Murray Hill, N. J.)

Mid-Amer. Elec. Conv., Hotel Muehlebach, Kansas City, Mo., Nov. 15–16. (DL*: June 15, J. Austin, Bendix Aviation Corp., 95 and Troost, Kansas City, Mo.)

PGPT Ann. Conf., Boston, Mass., Nov. 15–16. (DL*: June 1, C. W. Watt, Raytheon Co., Waltham, Mass.)

Symp. on Engineering Applications of Probability and Random Function Theory, Purdue University, Lafayette, Ind., Nov. 15–16.

1960 NEREM (Northeast Electronics Res. & Engrg. Mtg.), Boston, Mass., Nov. 15–17.

PGVC Ann. Mtg., Sheraton Hotel, Philadelphia, Pa., Dec. 1–2. (DL*: July 15, W. G. Chaney, American Telephone and Telegraph Co., 195 Broadway, N. Y. 7, N. Y.)

3rd EIA Conf. on Maintainability of Electronic Equipment, Hilton Hotel, San Antonio, Tex., Dec. 5–7.

1960 URSI-IRE Fall Meeting, Boulder, Labs., Boulder, Colo., Dec. 12–14.

Eastern Joint Computer Conf., New Yorker Hotel, New York, N. Y., Dec. 13–15. (DL, papers: Aug. 13, E. Kubie, Computer Usage Co., 18 E. 41 St., N. Y. 17, N. Y.)

1961

Symp. on Thermoelectric Energy Conversion, Statler Hotel, Dallas, Tex., Jan. 8–12. (DL*: Oct. 10, 1960, P. Klein, GE Co., Syracuse, N. Y.)

7th Natl. Symp. on Reliability and Quality Control, Bellevue-Stratford Hotel, Philadelphia, Pa., Jan. 9–11.

Symp. on Space Instrumentation, Washington, D. C., Jan. 16–17.

2nd PGMIL Conf., Biltmore Hotel, Los Angeles, Calif., Feb. 1–3. (DL*: Nov. 15, 1960, Dr. J. J. Meyers, Hoffman Electronic Corp., Military Products Div., 3717 S. Grand Ave., Los Angeles, Calif.)

* DL = Deadline for submitting abstracts.

GLOBECOM V Releases Call for Papers

The Fifth National Symposium on Global Communications, to be known as GLOBECOM V, will be held May 22–24, 1961, at the Hotel Sherman, Chicago, Ill. The conference will be sponsored by the American Institute of Electrical Engineers and the Professional Group on Communications Systems of the Institute of Radio Engineers.

Technical papers emphasizing the systems aspects of communications technology and related problems are solicited. Some of the areas of interest include:

Transmission Systems	Techniques & Problems	Systems Aspects
Active and Passive Satellites	Digital Data Handling	Planning
Cables	Propagation	Economics
Forward Scatter	Modulation	Switching
Microwave Relay	Multiplexing	Trunking
Long Haul Trunks	Interference	Frequencies

Abstracts of approximately 250 words, with a brief professional record of the author, should be submitted in duplicate by November 30, 1960, to: D. C. Campbell, Tech. Program Comm., I.T.T.—Kellogg, 5959 S. Harlem Ave., Chicago 38, Ill.

At the Symposium, oral presentation should require about 20 minutes, and 10 minutes will be allocated for discuss on. Due to the broad scope of subject matter, groups of authors are invited to submit related abstracts. The Technical Program Committee will endeavor to schedule all such parts, or make special provisions for a single presentation, should the material be of particular merit.

An abbreviated convention record will be prepared for use at the Symposium. Authors will be encouraged to submit their full length papers for later publication in the IRE Transactions on Communications Systems, or the AIEE bi-monthly *Communications and Electronics*.

Air Force MARS Announces November Schedule

The November schedule for the Air Force MARS Eastern Technical Net, operating Sundays from 2 to 4 P.M. EDT, on 3295, 7540, and 15,715 kc, has been announced as follows:

November 6 "Trip Report; Impressions of the Technological Status of Electronics Production in the U.S.S.R." E. R. Behn, Bosch-Arma Corp.

November 13 "Trip Report; Impressions of an American Engineer-Visitor to the U.S.S.R.," R. Popkin-Clurman, Telechrome Corp.

November 20 "Electromagnetic Compatibility," D. Wildfeuer, American Bosch-Arma Corp.

November 27 "Applications of Ultrasonics," A. Paley, American Bosch-Arma Corp.

December 4 "Principles of Guidance and Navigation, and Inertial Devices," D. P. Sarett, American Bosch-Arma Corp.

Calendar of Coming Events and Authors' Deadlines*

Int'l. Solid State Circuits Conf., University of Pennsylvania and Sheraton Hotel, Philadelphia, Pa., Feb. 15–17. (DL*: Oct. 14, 1960, J. J. Suran, Bldg. 3, Rm. 115, GE Co., Electronics Park, Syracuse, N. Y.)

Int'l. Symp. on Semiconductor Devices, UNESCO, 2 Place Fontenoy, Paris, Feb. 20–25.

Symp. on Engrg. Aspects of Magnetohydrodynamics, Univ. of Pa., University Park, Mar. 9–10.

IRE Int'l. Conv., N. Y. Coliseum and Waldorf-Astoria Hotel, New York, N. Y., Mar. 20–23.

PIB Int'l. Symp. on Electromagnetics and Fluid Dynamics of Gaseous Plasma, N. Y., N. Y., April 4–6. (DL*: Dec. 20, 1960, Symp. Committee, Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.)

SWIRECO, Dallas, Tex., April 19–21.

7th Region Tech. Conf. & Trade Show Westward Ho Hotel, Phoenix, Ariz., April 26–28.

Electronic Comp. Conf., Jack Tar Hotel, San Francisco, Calif., May 2–4.

5th Midwest Symp. on Circuit Theory, Allerton Park & Urbana Campus, Univ. of Ill., Urbana, May 7–8. (DL*: Oct. 1, Prof. M. E. VanValkenburg, Dept. EE, Univ. of Ill., Urbana.)

NAECON, Miami & Biltmore Hotels, Dayton, O., May 8–10.

Western Joint Computer Conf., Ambassador Hotel, Los Angeles, Calif., May 9–11. (DL*: Dec. 12, 1960, Dr. W. F. Bauer, Ramo-Wooldridge Co., 8433 Fallbrook Ave., Canoga Park, Calif.)

Microwave Theory and Tech. Nat'l. Symp., Sheraton Park Hotel, Washington, D. C., May 15–17.

GLOBCOM V, Sherman Hotel, Chicago, Ill., May 22–24. (DL*: Nov. 30, 1960, D. C. Campbell, I. T. T.-Kellogg, 5959 S. Harlem Ave., Chicago 38, Ill.)

Nat'l. Telemetering Conf., Chicago, Ill., May 22–24. (DL*: Dec. 15, 1960, J. Becker, AC Spark Plug Div., G.M. Corp., Milwaukee 1, Wis.)

5th Nat'l. Symp. on Product Engrg. and Production, Philadelphia, Pa., June 5–5.

5th Nat'l. Conv. on Military Electronics, Washington, D. C., June 26–28.

JACC, Univ. of Colorado, Boulder, June 28–30. (DL*: Abstracts, Nov. 15, 1960) Papers, Dec. 15, 1960, R. Kramer, 1961 JACC Program Committee, Bldg. 32, M.I.T., Cambridge, Mass.

4th Int'l. Conf. On Medical Electronics, Waldorf-Astoria Hotel, N. Y., N. Y., July 16–22. (DL*: April 1, 1961, H. P. Schwan, Moore School of E.E., Philadelphia 4, Pa).

WESCON, San Francisco, Calif., Aug. 22–25.

* DL = Deadline for submitting abstracts.

Call for Papers

1961 Joint Automatic Control Conference

University of Colorado, Boulder, June 28–30, 1961

A Call for Papers for presentation at the 1961 Joint Automatic Control Conference has been issued. Following the recent successful meeting in Cambridge, Mass., the five supporting societies—AIChE, AIEE, ASME, IRE, and ISA—announced that the 1961 JACC would be held at the University of Colorado on June 28–30, 1961.

Papers on any phase of automatic control work are being solicited. The subject matter may range from theory to applications and from systems to components. Papers should be submitted to: R. Kramer, IRE Representative, 1961 JACC Program Committee, Bldg. 32, M. I. T., Cambridge 39, Mass.

A one or two page abstract (four copies) must be submitted by Nov. 15, 1960. The full paper (four copies) suitable for review must be submitted by December 15, 1960.

1. Microscopic and Macroscopic Theories of Non-equilibrium Magneto-Fluidynamics.
   a) Distribution function vs moment approaches (cut-off procedures)
   b) Elastic and inelastic collision contributions
   c) Effects of heat conductivity, viscosity, etc.
   d) Effects of particle trapping, instabilities
2. Linear Wave Propagation in Uniform and Non-uniform Plasmas
   a) Classification of wave types (harmonic and other small disturbances)
   b) Surface waves
   c) Effects of plasma discontinuities
   d) Plasma radiation (random and ordered)
   e) Experimental wave techniques
3. Nonlinear Wave Propagation in Uniform and Non-uniform Plasmas
   a) Shocks
   b) Dissipation and excitation phenomena
   c) Harmonic generation and parametric effects
   d) Boundary layer effects
   e) Experimental techniques
4. Possible Applications of Magneto-Fluid-dynamics
   a) Electromechanical forces in moving plasmas
   b) Plasma generators
   c) Plasma jets
   d) Confinement and stability

This year the symposium is being organized by the Department of Aerospace Engineering and the Microwave Research Institute of the Polytechnic Institute of Brooklyn in cooperation with the IRE and the Institute of the Aeronautical Sciences and with the co-sponsorship of the Air Force Office of Scientific Research, the Office of Naval Research, and the U. S. Army Signal Corps.

In keeping with tradition, the symposium will endeavor to serve the twofold purpose of providing both a review of the present state of research in the above plasma fields and a forum for discussion of recent outstanding advances of interest to engineers, mathematicians, and physicists involved in plasma research. The program will comprise both invited and a limited number of contributed papers and will conclude with a round-table discussion of some future trends in plasma research. Further information may be obtained from Professors M. Bloom and E. Levi, Co-Chairmen of the Symposium Committee.

The closing date for submission of papers or 200-word abstracts is December 20, 1960. Correspondence should be addressed to: Symposium Committee, Polytechnic Institute of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

Conference on Magnetism and Magnetic Materials To Meet in New York

The Sixth Annual Conference on Magnetism and Magnetic Materials will be held November 14–17, 1960, in the Hotel New Yorker. This year, the AIEE shares joint sponsorship with the American Institute of Physics. As before, the IRE, The Metallurgical Society of the AIME, and the Office of Naval Research are cooperating with the sponsors. The program will consist of 21 invited and 130 contributed papers. Discussion will be invited following the presentation of each paper.

The four-day program will consist of fifteen sessions covering the following topics: Spin Arrangements, Computer Devices, Nuclear Hyperfine Fields, Metallic Thin Films, Ferromagnetic Resonance, High Coercive Materials, Exchange Interactions and Nonlinear Microwave Processes, Magnetization Processes and Fine Particles, Anisotropy, Domain Walls and Domain Wall Motion, Metals and Alloys, Microwave Devices, Oxides, and Soft Magnetic Metals.

The invited speakers will include several distinguished foreign scientists, and the guest speaker at a banquet to be held on November 16 will be T. K. Glennan, Administrator of the NASA.

Further information can be obtained from L. R. Bickford, Jr., Local Committee Chairman, IBM Research Center, Box 218, Yorktown Heights, N. Y.

PIB Symposium Issues Call for Papers

The eleventh international Polytechnic Institute of Brooklyn Symposium, scheduled for April 4–6, 1961 in New York City, will be devoted to "Electromagnetics and Fluid Dynamics of Gaseous Plasma." The area of interest covers collective electromagnetic and fluid dynamic phenomena and their self-consistent analytical description in the temperature range characteristic of the transition between partially and fully ionized gaseous plasmas.

Typical of the subjects planned for discussion at the Symposium are:

7th National Symposium on Reliability and Quality Control

Bellevue Stratford Hotel, Philadelphia, Pa., January 9–11, 1961

An advance program has been released for the Seventh National Symposium on Reliability and Quality Control, to be held in the Bellevue-Stratford Hotel, Philadelphia, Pa., from January 9–11, 1961. The IRE, AIEE, ASQC, and the EIA are the sponsoring organizations.

The Symposium will open with a keynote address by H. L. Hoffman, Hoffman Electronics Corp., whose topic will be "Reliability—Whose Responsibility?" Major General L. I. Davis, USAF, Commander, Patrick Air Force Base, Air Force Missile Test Center, will make the banquet address at the Symposium reception and banquet to be held on Tuesday evening, January 10. His topic will be "Reliability and Quality Control in Space." The following awards will be presented at the banquet: The 1960 National Reliability Award, presented by W. T. Sumerlin, General Chairman of the Symposium; the 1960 IRE Professional Group on Reliability and Quality Control Award, presented by P. K. McElroy, Chairman, PGRQC; and the 1960–1961 ASQC Electronics Division Award, presented by G. H. Beckhast, Electronics Division Chairman.

Standby Papers which are here listed will be published in the Symposium Transactions, along with the papers given at the Symposium:

"Reliability—Taking the Time to Evaluate," *L. G. Rado, General Precision Inc.*

"Reliability Trade-Offs for Repairable Systems," *G. Nagy, Goodyear Aircraft Corp.*

"Design Techniques Evaluation for Reliability," *R. W. Warr, General Electric Co.*

"An Analog Approach to Reliable Transistor Circuits," *H. T. Gruber, Battelle Memorial Institute.*

"Optimal Checking Procedures," R. W. Barlow, *Sylvania Electric Products, Inc.*

"A Critical Examination of Reliability Engineering," *E. V. Bersinger, Radio Corporation of America.*

"The Use of the Beta Distribution in Reliability," *J. H. Kao, Cornell Univ.*

"The Method of Lagrange Multipliers in Reliability," *K. L. Wong, Hughes Aircraft Co.*

"Model Analysis—Non-Linear Estimation Techniques," *N. Rosner, IBM Corp.*

Advance registration applications must be received before January 2, 1961. The fee for IRE, AIEE, and ASQC members is $20.00; for all others, it is $22.00. For further information, and for additional copies of the Symposium program, contact: J. Goodman, Publicity Chairman, RCA Bldg. 1-2, Mail Stop 4, Camden 2, N. J.

Monday Morning, January 9

Registration

Session 1—Keynote Address

"Reliability—Whose Responsibility?" *H. L. Hoffman, Hoffman Electronics Corp.*

Session 2A—Statistical Techniques

Moderator: *W. R. Pabst, Jr., U. S. Navy, Bureau of Weapons.*

"Some Discrete Life Testing Distributions," *J. G. Magistad, Sandia Corp.*

"Reliability Test Plans," *M. H. Saltz, Hughes Aircraft Co.*

"Sampling Plans Based on the Weibull Distribution," *H. P. Goode, Cornell Univ.*

Session 2B—Training Session

Moderator: *L. J. Paddison, Sandia Corp.*

"Basic Statistical Concepts," *F. M. Gryna, Jr., The Martin Co.*

"Reliability Prediction," *G. T. Bird, ARINC Research Corp.*

"Military Specifications Affecting Reliability," *H. L. Wuerffel, RCA.*

Session 2C—Reliability and Quality Control Education

Moderator: *R. T. Weil, Manhattan College.*

"Reliability Education—Success or Failure," *D. Meister, Convair.*

"Training for Reliability and Quality Control," *G. O. Hawley, Sandia Corp.*

"Reliability Awareness Campaign," *H. Kimel, GE Co.*

Monday Afternoon

Session 3A—Statistical Techniques

Moderator: *G. R. Herd, Booz-Allen Applied Research, Inc.*

"Statistical Methods Applied to Semiconductors," *W. R. Clemens, Hughes Aircraft Co.*

"Sequential Test to Measure and Improve Reliability," *S. A. Bessler, Sylvania Electric Products, Inc.*

"System Reliability Confidence Limits," *Miss B. A. Wales, IBM Corp.*

"Reliability of Parallel Systems With Repair," *S. G. Kneale, AVCO Corp.*

Session 3B—Training Session

Moderator: *L. J. Paddison, Sandia Corp.*

"Reliability Testing of Electronic Equipment," *A. C. Floyd, Hoffman Electronics Corp.*

"Manufacturing's Contributions to Reliability," *T. A. Budne, Statistical Engineering Consultant.*

"Failure Reporting and Corrective Action Programs," *J. R. Holmes, IBM Corp.*

"Field Reliability," *H. Bayer, Douglas Aircraft Co.*

Session 3C—Reliability Management

Moderator: *Capt. R. E. Foster, U. S. Navy, Bureau of Ships.*

"Reliability from the Program Manager's Standpoint," *C. M. Ryerson, RCA.*

"Intra-Corporation Reliability Coordination," *W. J. Picker, General Precision, Inc.*

"Reliability Policies in a Multi-Product Company," *J. M. Wuerth, North American Aviation, Inc.*

"Case History of a Corporate Reliability Program," *F. McGinnis, Sperry Gyroscope Co.*

Monday Evening

Session 4—Symposium Round Table Discussion

Moderator: *C. M. Ryerson, RCA.*

Topics: "How Military and Industry can work together toward better reliability training"; "How incentives and penalties can be used in contracts to achieve higher equipment reliability"; "How 'small business' can play a part in helping procure more reliable equipment"; "The most important policy matters in reliability that require national attention during 1961."

Tuesday Morning, January 10

Registration

Session 5A—Models and Mathematics

Moderator: *Mrs. J. R. Rosenblatt, National Bureau of Standards.*

"A Unique Allocation of Required Part Reliability," *A. M. Breipohl, Sandia Corp.*

"Distributions and Their Properties," *A. J. Finocchi, IT&T Corp.*

"Failure Rate Study for Log-Normal Lifetime Model," *Mrs. L. R. Goldthwaite, Bell Telephone Labs.*

"Improving System Reliability," *K. M. Hall, Sylvania Electric Products, Inc.*

Session 5B—Accelerated Testing

Moderator: *K. W. Davidson, Texas Instruments, Inc.*

"Accelerated Life Testing of Mica Capacitors," *H. S. Endicott, GE Co.*

"Accelerated Reliability Testing of Equipment," *A. D. Pettinato, Rome Air Dev. Ctr.*

"Accelerated Tests and Predicted Capacitor Life," *D. Mannheim, Sprague Electric Co.*

"Stress Aging to Failure of Semiconductor Devices," *G. A. Dodson, Bell Telephone Labs.*

Session 6A—System Reliability

Moderator: *Col. W. Scandrett, Office Chief Research and Development, U. S. Army.*

"Designing Reliability in the B-58 BNS," *M. D. Katz, Sperry Rand Corp.*

"Use and Repair Effects on Availability/Reliability," *P. R. Taylor, IBM Corp.*

"Redundancy Considerations in Satellite Systems," *R. H. Myers, Hughes Aircraft Co.; L. A. Aroian, Space Technology Labs.*

"Effects of On-Off Cycling on Equipment Reliability," *R. J. Boteilho, ARINC Research Corp.*

Tuesday Afternoon

Session 6B—Cost Considerations

Moderator: *R. H. DeWitt, Office of Electronics, ODDR&E.*

"Minimizing the Cost of Reliability Testing," *R. T. Maloney, Sperry Rand Corp.*

"AF Production Experience on AGREE Procedures," *G. W. Lindsay, Air Materiel Command.*

"The Dollar Value of Improved Part Reliability," *E. L. Welker, ARINC Research Corp.*

"Design Goal for Reliability," *I. Bosinoff, Sylvania Electric Products, Inc.*

Wednesday Morning, January 11

Registration

Session 8A—Design Techniques

Moderator: *Lt. Col. T. T. Brundage, Air Research and Development Command.*

"Design Review—The Why and the How," *R. M. Jacobs, Sylvania Electric Products, Inc.*

"Self Checking in Large Scale Digital Computers," *H. S. Rasmussen, IBM Corp.*

"Digital Computers for Circuit Analysis," *H. S. Scheffler, Battelle Memorial Inst.*

"The Road to Reliable Design," *R. T. Williams, ARINC Research Corp.*

Session 8B—Reliability Prediction

Moderator: *N. E. Golovin, Rabinow Engrg. Co., Inc.*

"Reliability Growth Prediction During Design," *D. R. Earles, The Martin Co.*

"Reliability Measurability of a Costly Missile," *E. Pieruschka, Lockheed Aircraft Corp.*

"Predicting Failures in a Guidance Set," *W. T. Wells, Research Triangle Inst.*

"Transistor Reliability Philosophy Applied to BMEWS," *E. M. Yanis, RCA.*

Wednesday Afternoon

Session 9A—Reliability and Quality Control Programs

Moderator: *A. M. Okun, RCA.*

"Quality Control in Research and Development," *R. G. Fitzgibbons, Raytheon Co.*

"Case History of Product Assurance on Polaris," *Cmdr. R. A. Weatherup, U. S. Navy; W. Huber, Sperry Rand Corp.*

"Results of an Operational Reliability Program," *M. A. Anderson, Hughes Aircraft Co.*

Session 9B—Reliability Testing

Moderator: *L. J. Jacobson, International Resistance Co.*

"Methods of Determining Circuit Reliability," *D. M. Larson, IBM Corp.*

"The Achievement of Reliability in a Switch Program," *R. D. Murdoch, Sandia Corp.*

"Reliability Testing of Aircraft Instruments," *A. C. Evans, Daystrom, Inc.*

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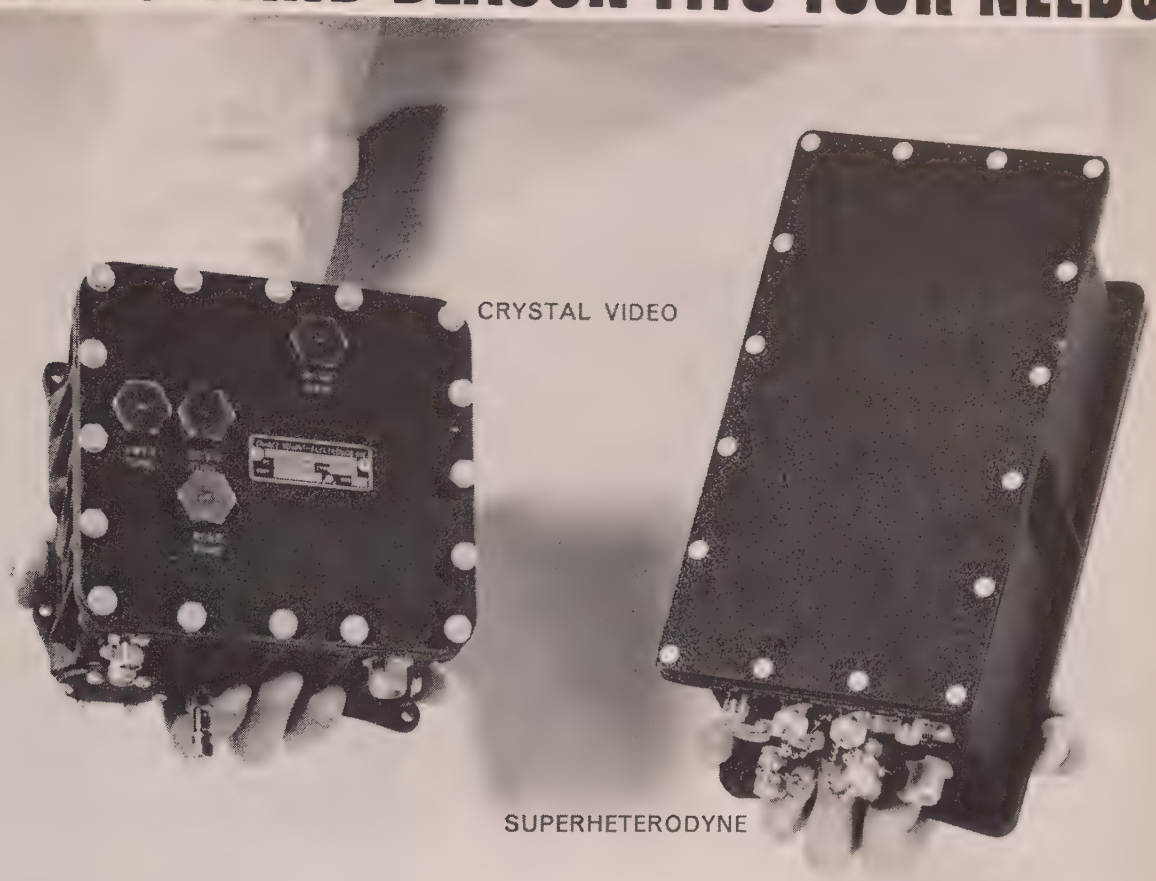

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Flight tested, off-the-shelf delivery. Both beacons have been flight tested and successfully operated at theoretical line-of-sight ranges to 60,000 feet. Lightweight, compact and rugged enough to operate under extreme field conditions, the aluminum alloy casting packages are pressurized, designed for hard mounting to the airframe. Accessible parts and readily available adjustments insure easy service. Both beacons are completely transistorized throughout, except for the magnetron and local oscillator.

Compatibility Vought Electronics beacons are designed to be compatible with existing C-Band radars (FPS-16, MPS-26, etc.) and with primary power and antennas available on modern aircraft.

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1960 Eastern Joint Computer Conference

New Yorker Hotel, New York, N. Y., December 13–15

The program of technical papers has been announced for the 1960 Eastern Joint Computer Conference, which will be held in New York this year, on December 13–15. The IRE, AIEE, and Association for Computing Machinery are co-sponsors.

Comprehensive coverage of progress and applications of the computer field will be provided by speakers from government, military, industry, and research facilities.

Tuesday Morning, December 13

Opening Remarks: *N. Rochester, Conference Chairman, IBM, and E. C. Kubie, Program Chairman, CUC.*

"A Logical Machine for Measuring Problem Solving Ability," *C. R. Langmuir, Psychological Corp.*

"A Method of Voice Communication With a Digital Computer," *S. R. Petrick and H. M. Willett, AF Cambridge Res. Labs.*

"FILTER—A Topological Pattern Separation Computer Program," *D. Innes, Lawrence Radiation Lab.*

"Redundancy Exploitation in the Computer Solution of Double-Crostics," *E. S. Spiegelthal, Consultant.*

Tuesday Afternoon

"A Computer for Weather Data Acquisition," *P. Meissner, J. Cunningham, and C. Kettering, Natl. Bur. of Standards.*

"A Survey of Digital Methods for Radar Data Processing," *F. H. Krantz and W. D. Murray, Burroughs Corp.*

"The Organization and Program of the BMEWS Checkout Data Processor," *A. Eugene Miller, Auerbach Electronics Corp., and M. Goldman, RCA.*

"Ultra-High Speed Dynamic Display System for Digital Data," *B. G. Tregub, Melpar.*

"High Speed Data Transmission Systems," *R. G. Matteson and J. D. Barnard, Stromberg-Carlson Co.*

Wednesday Morning, December 14

"Parallel Computing With Vertical Data," *W. Shooman, System Development Corp.*

"The TABSOL Concept," *T. F. Kavanagh, General Electric Co.*

"Theory of Files," *L. Lombardi, University of California.*

"Polyphase Merge Sorting—An Advanced Technique," *R. L. Gilstad, Minneapolis-Honeywell Regulator Co.*

"The Use Of Binary Computer For Data Processing," *G. H. Redmond and D. E. Mulvihill, Chrysler Corp.*

Wednesday Afternoon

"High Speed Printer and Plotter," *F. T. Innes, Briggs Associates, Inc.*

"A Description of the IBM 7074 System," *R. R. Bender, D. T. Doody, and P. N. Stoughton, IBM, Poughkeepsie.*

"The RCA 601 System," *D. L. Nettleton and K. K. Kozarsky, RCA.*

"Associative Self-Sorting Memory," *R. R. Seeber, Jr., IBM, Poughkeepsie.*

"UNIVAC—RANDEX II—Random Access Data Storage System," *G. J. Axel, Remington Rand.*

"Hot-Wire Anemometer Paper Tape Reader," *J. H. Jory, Soroban Engineering, Inc.*

Thursday Morning, December 15

"Data Processing Techniques in Design Automation," *W. L. Gordon, Minneapolis-Honeywell Regulator Co.*

"Impact of Automation on Digital Computer Design," *W. A. Hannig and T. L. Mayes, General Electric Co.*

"Calculated Waveforms For Tunnel Diode Locked Pair," *H. R. Kaupp and D. R. Crosby, RCA.*

"On Iterative Factorization in Network Analysis by Digital Computer," *W. H. Kim, C. V. Freiman, and W. Mayeda, Columbia University.*

"A Computer-Controlled Dynamic Servo Test System," *V. A. Kaiser and J. L. Whittaker, Douglas Aircraft Co.*

Thursday Afternoon

"The Flying Spot Scanner As An Input Sensor To A Character Reading System," *J. S. Bryan, J. B. Chatten, F. P. Keiper, and C. F. Teacher, Philco Corp.*

"Use of A Digital/Analog Arithmetic Unit Within A Digital Computer," *D. Wortzman, IBM.*

"PB-250 A High Speed Serial General Purpose Digital Computer Using Magnetostrictive Delay Line Storage," *R. M. Beck, Packard Bell Computer Corp.*

"The Instruction Unit of the Stretch Computer," *R. T. Blosk, IBM, Poughkeepsie.*

"The Printed Motor: A New Approach To Intermittent and Continuous Motion Devices in Data Processing Equipment," *R. P. Burr, Circuit Research Co.*

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Sheraton Hotel, Philadelphia, Pa., December, 1–2, 1960

"New Horizons in Vehicular Communications" will be the theme of the two-day annual conference of the IRE Professional Group on Vehicular Communications. Papers will stress new components, circuits, applications and measurement techniques.

Sir Robert Watson Watt, noted scientist and radar inventor, will be guest speaker at a dinner on December 1, and Dr. George H. Brown, Vice President, RCA, will address a luncheon audience the following day.

Conference information and reservations are available from Marion T. Speights, Registration Chairman, P.O. Box 443, Fort Washington, Pa.

A partial listing of papers has been announced as follows.

"A New Two-Way Dial Mobile Radio System for Telephone Construction and Maintenance," *Rice and Flaminio, Bell Telephone Co. of Pennsylvania.*

"The Citizen Band and Boom," *I. Loucks, General Land Transportation Div., FCC.*

"A Proposed System for 11-Channel Access in the Mobile Telephone Service," *C. A. Rypinski, Secode Electrical Communications.*

"Speech Modulation and Limiting," *L. Getgen, Lenkurt Co.*

"Transmission Line Reliability," *L. A. Bondon, Prodelin.*

"Mechanical Filter for FM Mobile Application," *R. A. Johnson, Collins Radio Co.*

"Mobile Integrating Circuit," *H. H. Davids, General Electric Co.*

"Sinad System Design," *R. A. Farmer, General Electric Co.*

"Personal Radio Systems and Applications," *J. F. Mitchell, Motorola, Inc.*

"Measurement of Deviation and Frequency in Two-Way Radio Systems," *M. Cooper, Motorola, Inc.*

"A Modular Approach to the Design of FM Communications Equipment," *B. Tennent and G. G. Armitage, Ferritronics Ltd.*

"A New Helical Ground Plane Antenna for 30 to 50 Mc," *L. H. Hansen, Andrew Corp.*

"A Centralized Mobile Telephone Test Panel," *J. P. Gill.*

"The Application of Telephone Trunking Principals to Multichannel Mobile Service," *E. S. Randel, AT&T Co.*

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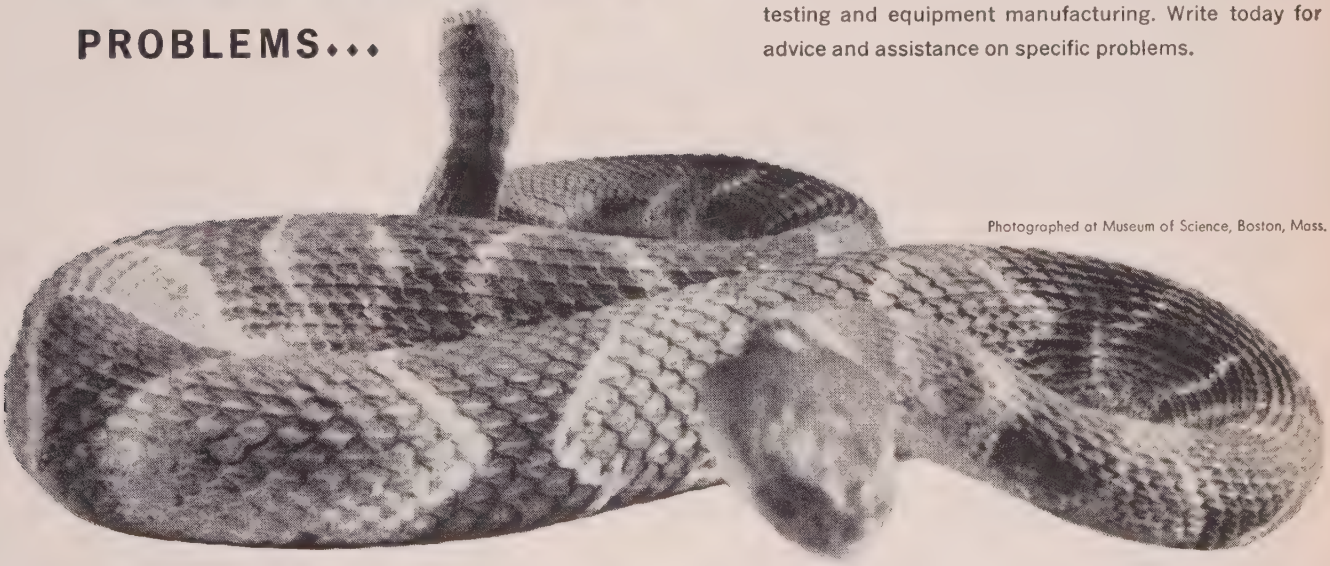

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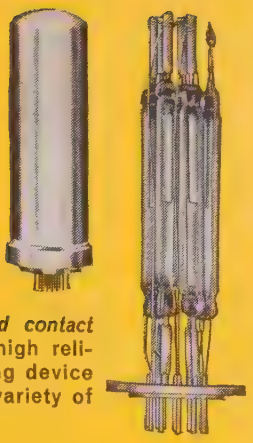

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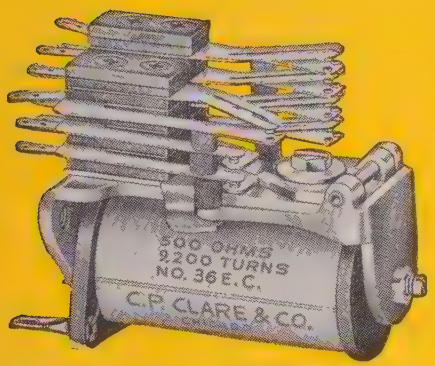

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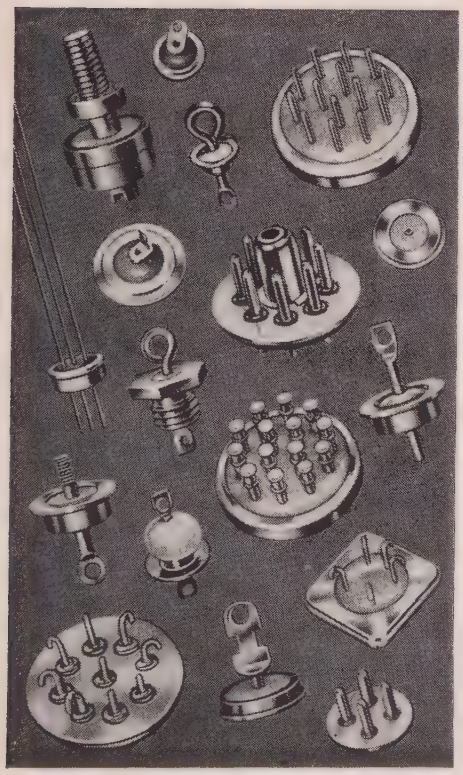

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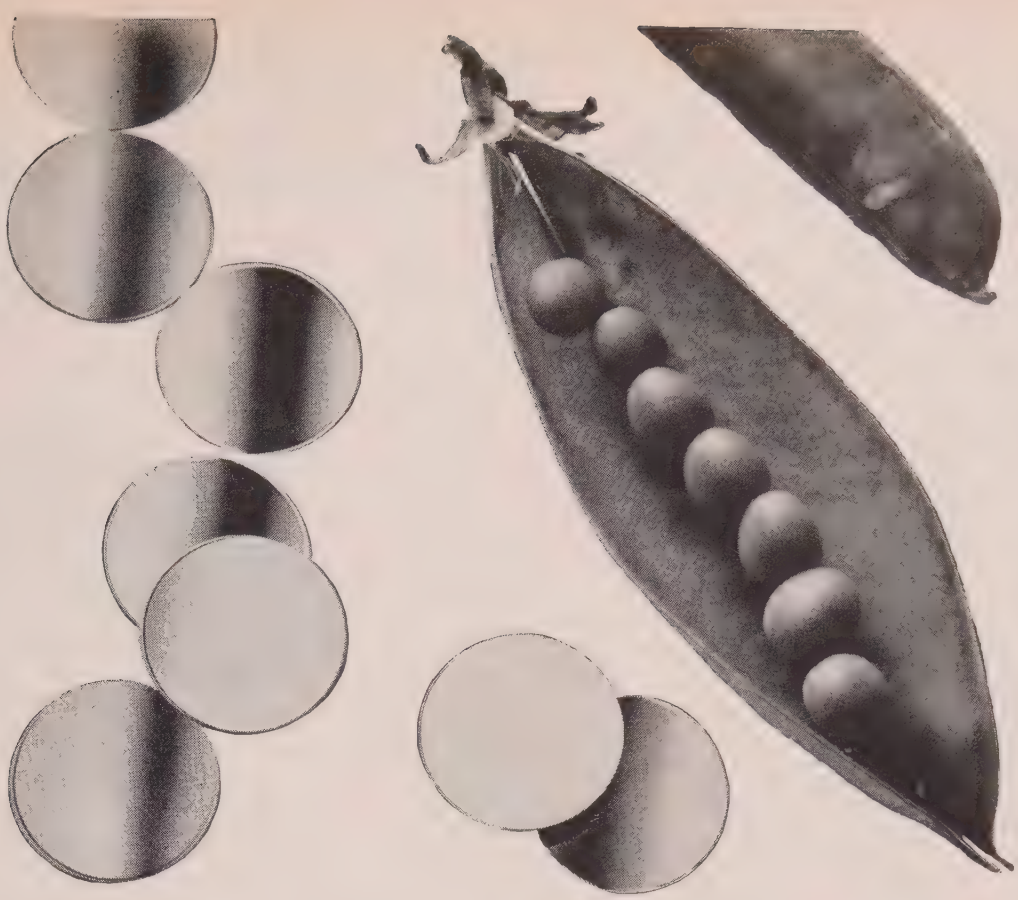

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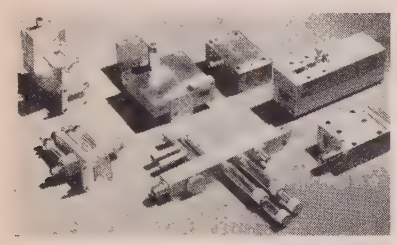
NEWS
New Products

Parametric Amplifiers

Availability of custom-designed, reactance diode parametric amplifiers for applications at all microwave frequencies was announced by the **Apparatus Div., Texas Instrument Incorporated,** 6000 Lemmon Ave., Dallas 9, Texas.



Simultaneous TI research programs on materials, device development and circuits have produced outstanding gain, bandwidth, and noise figures in amplifiers ranging from L through K_u bands, the firm said. These devices are claimed to make it possible to obtain lower noise figures and subsequently greater range in radar, telemetry, communications and a host of other electronic systems.

The heart of the parametric amplifier line is the gallium arsenide, variable reactance (varactor) diode (produced by TI's Semiconductor-Components division) which gives greatly enhanced noise performance over a wide range of temperatures down to liquid nitrogen temperature (77°K). This material is being used because TI experience has showed it has higher cut-off frequencies with correspondingly lower noise figures and higher frequency operation than any material thus far investigated.

Amplifiers available from TI include up-converters, and non-degenerate and degenerate negative resistance amplifiers. Any of these may have certain advantages which make it best for a given application, and TI advises that it can provide the assistance necessary to determine the type amplifier, noise figure, gain and bandwidth needed for a given requirement.

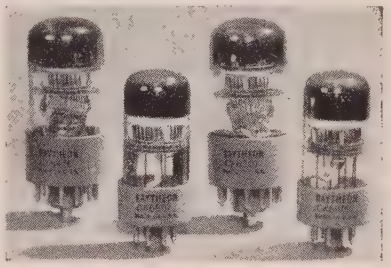
Typical amplifiers developed by TI are: an S-band, Model S-21, tunable from 2.7 to 2.9 Gc, which has conservative ratings of 3 db overall noise figure (double sideband) with 15 db gain over a 15 mc bandwidth; Model C-11, a c-band unit, which is tunable from 4.4 to 5.0 Gc and has a 3 db overall noise figure (double sideband) with 15 db gain over a 20 mc bandwidth; and L-band unit, Model L-11, which is tunable from 0.9 to 1.2 Gc and offers a 1 db noise figure (double sideband) with 16 db gain over a 50 mc bandwidth. Other models in late stages of development include X- and K_u-band units which are yielding

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

usually high gain/bandwidth characteristics with low noise figures.

Decade Counters

Electron tubes that count up to 100,000 units per second and visually display the total on their faces, were introduced by **Raytheon Company's Industrial Components Div.,** 55 Chapel St., Newton, Mass.



Called decade counters, the tubes record and indicate the count on ten glowing cathodes in luminescent dials within the tubes. With numbered faceplates, placed on series-connected decade counters, the total can be read visually in units, tens, hundreds, etc., during operation.

Applications engineers foresee the decade counter's use in assembly-line packaging, in measuring pipeline flow and in programming elapsed-time intervals, such as sequency welding.

Expected to play a special role in nucleonics, the decade counter can aid in plotting the spectrum of isotopes over a time-period or the energy level of radiation impinging on such nuclear energy detecting devices as scintillometers.

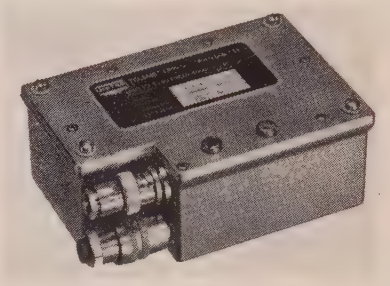
Designed for use in automatic machine tool programming and control, and for high-speed processes such as coil and transformer winding, the tube's count is converted to electrical pulses for readout.

The CK6909 and CK6910 decade counters operate up to 100 kc and the CK6802 and CK6476 operate up to 4 kc. The tubes are a cold cathode, bidirectional, ring-stepping device featuring low dissipation and short resolution time.

The CK6476 and CK6910 offer electrical readout from any cathode lead while the CK6802 and the CK6909 offer electrical readout at the 0, 5, 8 and 9 cathode leads.

Additional information on the decade counters may be obtained from William Weed at the firm.

Telemetry Transmitter



ElectroDynamics, Inc., 200 Allendale Rd., Pasadena, Calif., has developed the Model TR-10, a transistorized FM telemetry transmitter for the company's FM/FM and PCM systems and flight-proved in numerous military applications. TR-10 is scheduled for use in such missiles as Minuteman, Terrier, Javelin/Journeyman, and Hound Dog and in space programs which include Midas, Samos and Pioneer. The transmitter is now entering the market, available to needs of other industry and services. Specifications of the TR-10 include: 2.5 watts output with true FM modulation over complete 215–265 mcs telemetry band; 99.9% reliability for 500 hours; modulation frequency response ±2 db from 3 cps to 300 kc; vibration-induced noise less than 1.5 kc deviation at 20 G's from 20 cps to 2000 cps; modulation linearity less than one percent from straight line at 125 kc deviation. The TR-10 qualifies to radio noise specification MIL-STD-442, is hermetically sealed and exceeds military environmental specifications including MIL-E-5272. Data sheets available from the firm.

Switching Triode

A new high-voltage triode that can switch 150,000 volts in pulse modulators for radars and other electronic switching applications has been announced by **The Machlett Laboratories, Inc.,** 1063 Hope St., Springdale, Conn., a subsidiary of Raytheon Co.

The oil-cooled ML-7668 has dc plate and inverse voltages of 150,000 volts and a peak cathode current of 6.5 amperes. Dissipation of grid and plate are 50 and 750 watts, respectively. At 150,000 volts the cut-off grid voltage is in the order of only −900 volts.

Filament voltage is 12.6 volts and filament current is approximately 29 amperes. The new triode measures $11\frac{3}{8}$" high with a $5\frac{1}{8}$" diameter.

Full specifications for the ML-7668 are available from the firm.

(Continued on page 148A)

Creative Microwave Technology

Published by MICROWAVE AND POWER TUBE DIVISION, RAYTHEON COMPANY, WALTHAM 54, MASS., Vol. 2, No. 5

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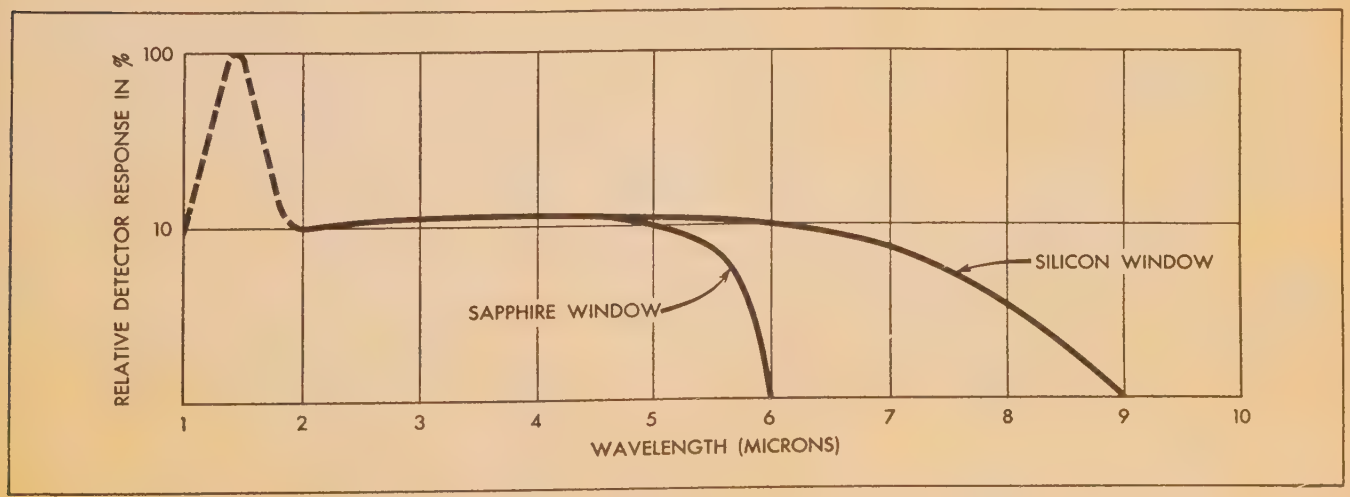

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- ISOLATION: 15 to 30 db depending upon range
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FXR's Ferrite Isolators are broadband, high performance waveguide and coaxial microwave components which provide maximum isolation and minimum insertion loss. In general these isolators are used in any application where it is desired to attenuate either the forward or reverse power flow without corresponding attenuation in the opposite direction. They are used to reduce the VSWR presented by a load and to isolate the oscillator for more stable operation.

Model No.	Frequency Range KMc	Minimum Isolation db	Peak Power	Price
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C157A	5.85- 8.20	20	*2KW	245.00
W157A	7.05-10	24	*2KW	245.00
X157A	8.20-12.4	30	*2KW	220.00
Y157A	12.40-18	24	**1KW	245.00
K157A	18.00-26.50	24	**1KW	270.00
K157AF†				270.00
N157A	2.00- 4.00	15	**2KW	450.00

*Load VSWR 5 **Load VSWR 2
†K157AF has the same specifications as K157A, except for the flange.

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November 15-17, 1960

For your convenience in planning your visit to NEREM, we are listing below the exhibitors who will appear at this show. A complete technical program for the NEREM show appeared on pages 35A-41A in the October 1960 issue.

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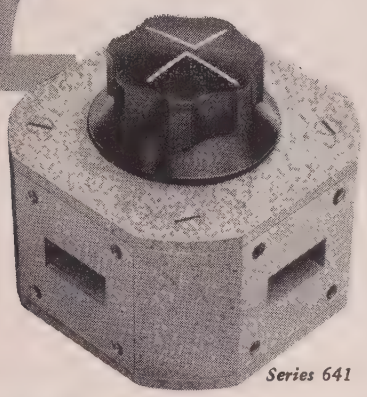
2



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- RANGE: 3.95 KMc to 40.00 KMc (in 7 sizes)
- CROSSTALK: 60 db min
- VSWR: 1.10 max
- Choice of manual or electrical drive
- High-power capacity

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MODEL NO.	FREQUENCY RANGE KMc	WAVEGUIDE TYPE RG-( )/U	*PRICE (MANUAL)
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C641A	5.85- 8.20	50	300.00
W641A	7.05-10.00	51	265.00
X641A	8.20-12.40	52	225.00
Y641A	12.40-18.00	91	250.00
K641A, AF	18.00-26.50	53	275.00
U641A, AF	26.50-40.00	96	300.00

*Slightly higher for electrically driven units.

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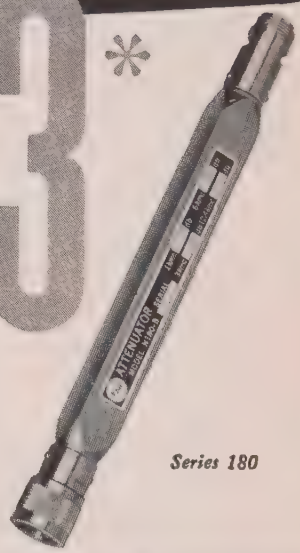
FXR's BROADBAND FIXED COAXIAL ATTENUATORS

3



Series 180

- FREQUENCY RANGE: 0.6 KMc to 12.4 KMc
- ATTENUATION VALUES: 3, 6, 10, 20 db
- CONNECTORS: Type N — male one end, female the opposite end

No. 3 of a series of FXR's new precision microwave components designed to meet the ever-growing needs of the microwave industry.

FXR's Broadband Fixed Coaxial Attenuators are extremely useful and completely dependable in applications requiring isolation between RF components and extending power meter ranges. They may also be used for the calibration of directional couplers, in obtaining antenna characteristics and for similar applications. These attenuators have exceptional stability and are capable of withstanding appreciable overloads and peak power with no change in characteristics. They have high shock and vibration resistance and exhibit a negligible change of attenuation under humidity and temperature cycling.

Model No.	Frequency KMc	Max. VSWR	Frequency Sensitivity db	Price
N180A	.6-11.0	1.3	(− .3) (+ .5)	$42.00
N180B	1.0-11.0	1.3	(− .6) (+ .7)	42.00
N180C	1.0- 2.0 2.0-11.0	1.35 1.30	(−1.2) (+1.3)	42.00
N180D	2.0- 3.0 3.0-11.0 11.0-12.4	1.35 1.30 1.40	(−1.3) (+1.9)	42.00

Write for Catalog Sheet No. 180

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Design-Development-Manufacture
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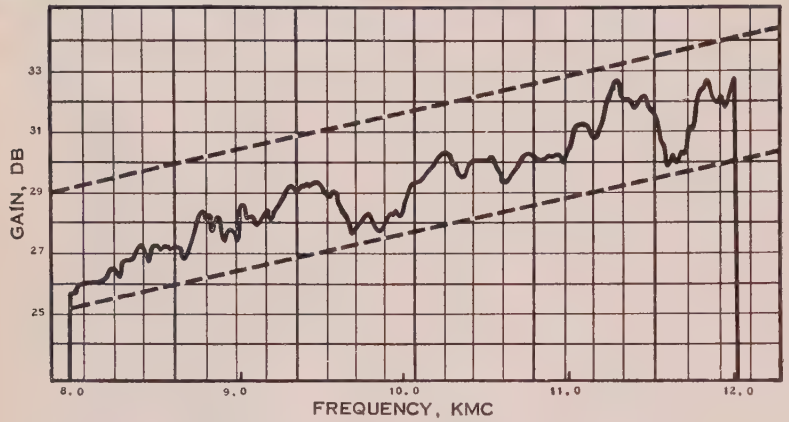
Woodside, N. Y./TWX: NY 43745

SLOPE GAIN TWT FOR IMPROVED SYSTEM OPERATION



RECORDED PLOT OF SMALL SIGNAL GAIN VS FREQUENCY
OF A MODIFIED HUGGINS HA–20 PM FOCUSED TWT AMPLIFIER

The production department at Huggins Laboratories has become very adept at providing traveling wave tubes having specific performance characteristics. These characteristics generally have stressed achievement of a prescribed small signal gain as a function of frequency over definite frequency bands, depending on customer requirements.

As an example, tubes can be provided in which small signal gain varies at some prescribed rate as a function of frequency. The use of a TWT whose gain increases as frequency increases makes it possible to compensate for losses of other microwave system components, which generally increase with frequency, also. The over-all result is a system which, between two given points, has a response which is very nearly independent of frequency. Traveling wave tubes having such properties have been supplied over several specific frequency ranges within the 2.0 to 12.4 KMC bands.

The curve above gives an example of the extent to which the small signal gain response of a TWT amplifier may be controlled. Here, the modification of a standard X-band PM-focused amplifier resulted in an average gain which increased by 1.5 db per 1000 MC increase in frequency over the 8.0 to 12.0 KMC band. Response of this type is possible with no adjustment necessary by the user external to the tube — the curve is presented with all potentials and currents fixed. Other types of gain responses are also possible, such as TWT amplifiers whose gain varies at some fixed rate over certain particular frequency bands.

The curve is a plot made with a pen recorder used in conjunction with a constant power system. This system makes use of a gridded low-level TWT and the use of feedback to control its output such that it is very nearly constant as a function of frequency and drive (over certain input level ranges). Such a system is described in Huggins Engineering Note, Number 8, *"The use of the TWT in constant power systems."*

A copy of this is available upon your request, and is bound in our two-volume catalog set which is also available should you not already be on our mailing list. Submit inquiry on company letterhead.

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A-27
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LOW-LOSS RF LACQUER

• Q-Max, an extremely low loss dielectric impregnating and coating composition, is formulated specifically for application to VHF and UHF components. It penetrates deeply, seals out moisture, provides a surface finish, imparts rigidity and promotes stability of the electrical constants of high frequency circuits. Its effect upon the "Q" of RF windings is practically negligible.

• Q-Max applies easily by dipping or brushing, dries quickly, adheres well, meets most temperature requirements. Q-Max is industry's standard RF lacquer. Engineers who know specify Q-Max! Write for new catalog.

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Communication Products Company, Inc.

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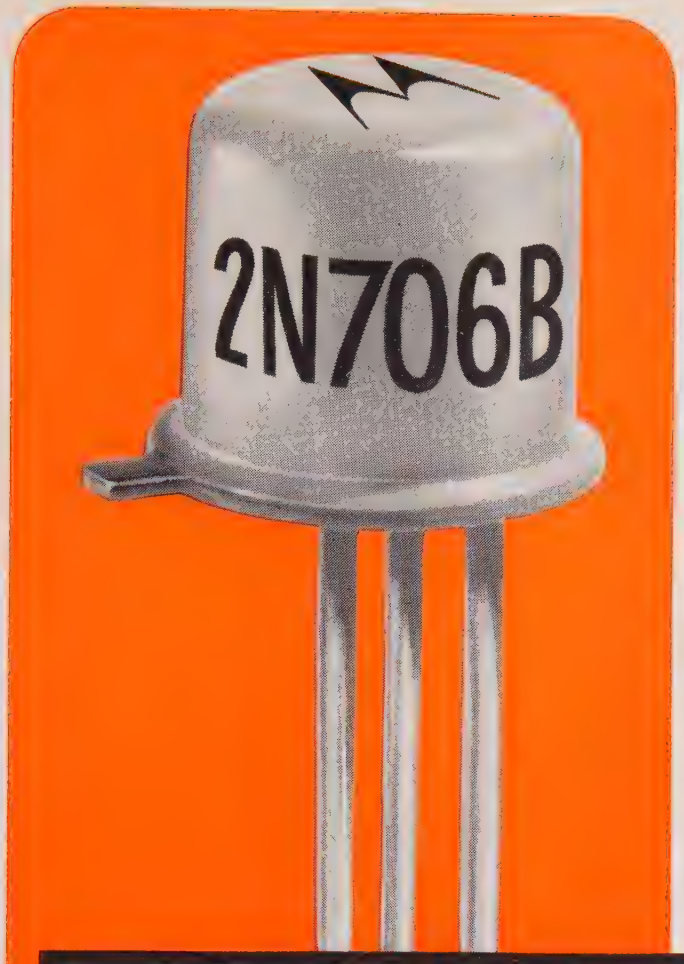

now a low storage silicon mesa with the high reliability... high uniformity of the Motorola germanium mesa

ANNOUNCING MOTOROLA'S 2N706B SILICON MESA

The new Motorola 2N706B high-speed switching transistor has all the advantages of 2N706 and 2N706A... with improved high-frequency performance, lower base resistance (r'_b) and lower saturation voltage ($V_{CE(SAT)}$).

NOW ALL THREE **2N706B** - 2N706A - 2N706 FROM MOTOROLA

ELECTRICAL CHARACTERISTICS at 25°C

PARAMETER	SYMBOL	2N706B	2N706A	2N706	UNIT
Collector-Base Voltage	BV_{CBO}	25	25	25	volts
Collector-Emitter Voltage	BV_{CER}	20	20	20	volts
Emitter-Base Voltage	BV_{EBO}	5	5	3	volts
Power Dissipation @ 25°C case temperature	P_D	1.0	1.0	1.0	watts
Power Dissipation @ 25°C ambient temperature	P_D	0.3	0.3	0.3	watts
Forward Current Transfer Ratio	h_{FE}	20-60	20-60	20 min	—
Base-Emitter Saturation Voltage	$V_{BE(SAT)}$	0.7-0.9	0.7-0.9	0.9	volts
Collector Saturation Voltage	$V_{CE(SAT)}$	(0.4)	0.6	0.6	volts
Base Resistance	r'_b	(50)	—	—	ohms
Storage Time	t_s	25	25	60	nsec
Turn On Time	t_{on}	40	40	—	nsec
Turn Off Time	t_{off}	75	75	—	nsec

Actual Size

The new 2N706B Motorola silicon mesa transistor is available now for your critical aircraft, missile, industrial and computer circuitry. In addition, Motorola is also supplying the 2N706A and 2N706. These Motorola silicon devices utilize a new, exclusive "E structure" mesa configuration which provides the same high uniformity and extreme reliability that has made Motorola the world's largest producer of high-quality germanium mesas.

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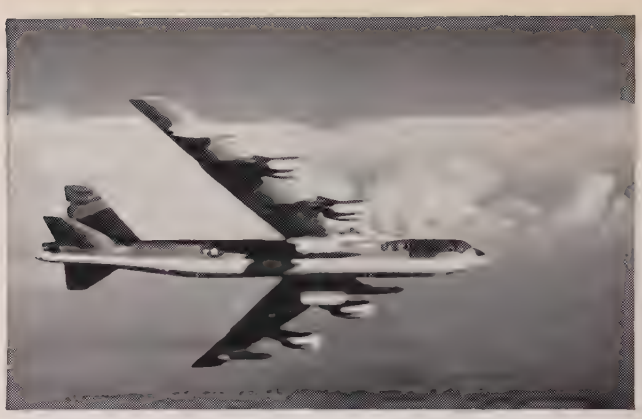
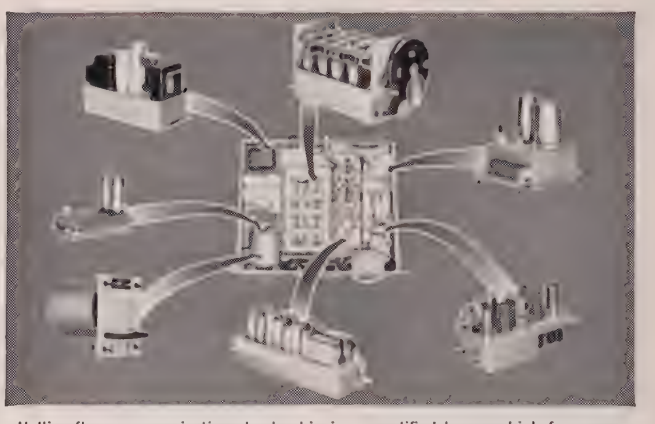
For more than a quarter century, Hallicrafters has worked in close partnership with our armed forces on fast solutions to critical military electronics problems. Out of this priceless experience are emerging startling new ideas and hard-hitting, fast-moving techniques to keep our country one jump ahead in electronic warfare . . .



B-52 and other military aircraft will be protected by the most potent Electronic Countermeasures equipments yet devised. These equipments were developed in close teamwork with the Air Force under Hallicrafters' QRC (Quick Reaction Capability) program. Now qualified to meet full environmental specifications, they are in quantity production.

New levels of speed and efficiency are being reached in equipment modernization, retrofit and technical support programs with Hallicrafters' radical new "Blue Streak" project. Specially-trained Maintenance and Technical Support Teams, close-knit and flexible, can be tactically deployed to accomplish maintenance, installation and testing of electronics weapons systems anywhere in the world.



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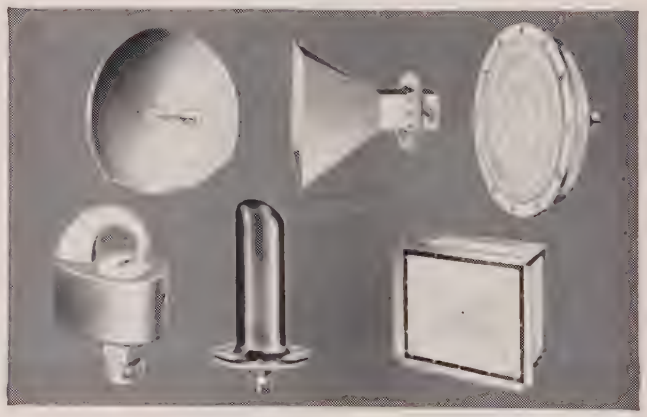
Hallicrafters participation in the Atlas missile project helped to develop capability for many areas of the complex missile field, including code translator data systems; ground support equipment; ECM testing and antenna systems. Current explorations involve latest Infra Red techniques.



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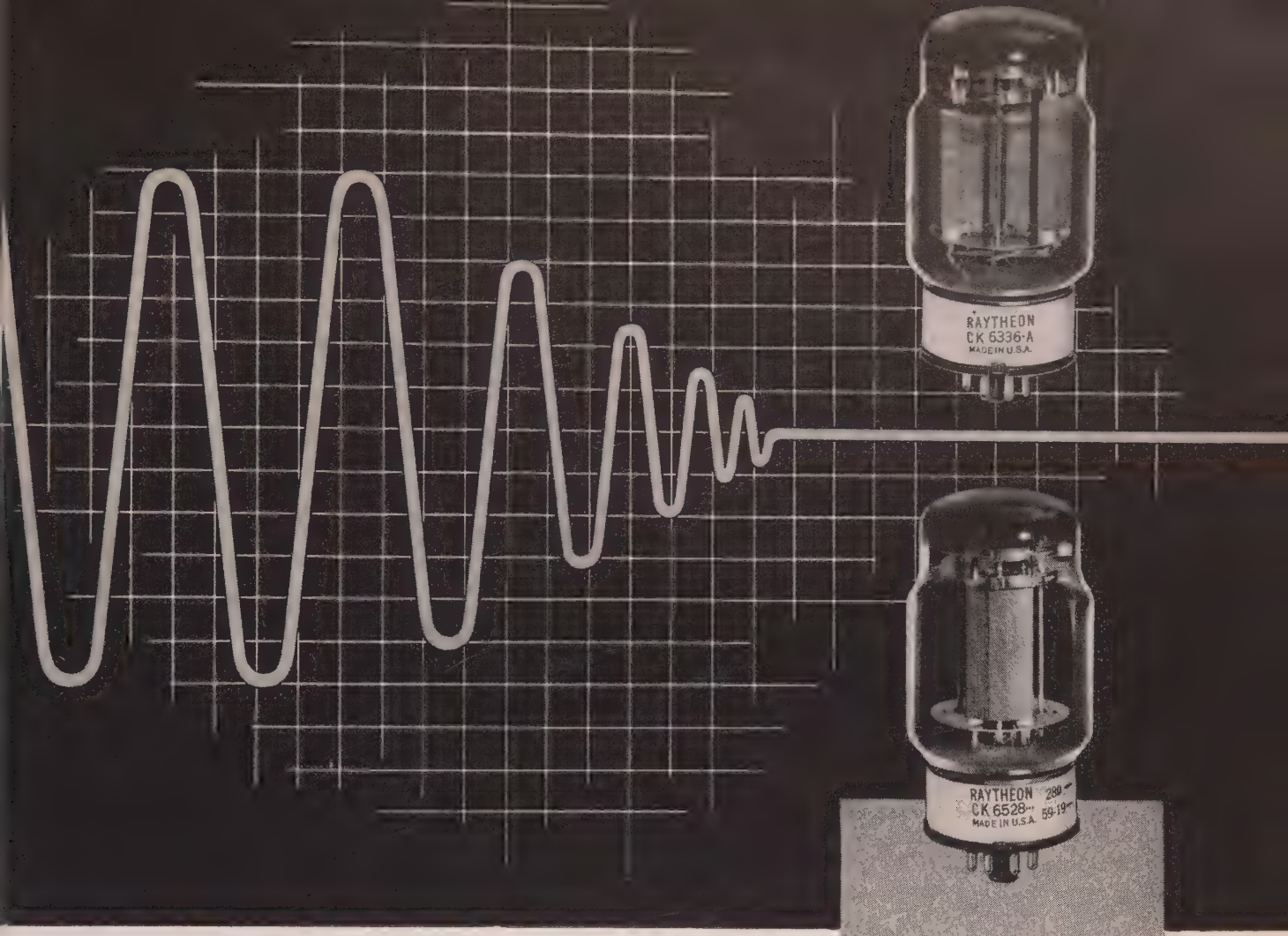

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RATINGS

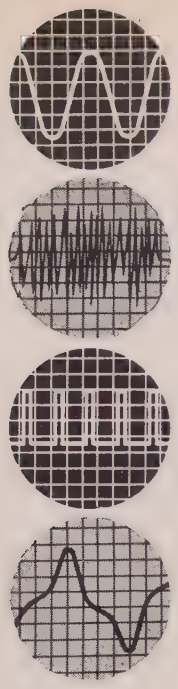
	Max. Plate Voltage	Max. Plate Dissipation Watts	Max. Plate Current (per plate)	Amplification Factor
CK6336A	400	2 x 30	400 mAdc	2.7
CK6528	400	2 x 30	300 mAdc	9

RAYTHEON CK6336A AND CK6528

RAYTHEON COMPANY

INDUSTRIAL COMPONENTS DIVISION

RAYTHEON



TRUE RMS

frequency range 5 to 500,000 cps

FEATURES

Built-in calibrator . . . easy-to-read 5 inch log meter . . . immunity to severe overload . . . useful auxiliary functions

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ACCURACY: 3% from 15 cps to 150KC; 5% elsewhere. Figures apply to *all* meter readings

MAXIMUM CREST FACTORS: 5 at full scale; 15 at bottom scale

CALIBRATOR STABILITY: 0.5% for line variation 105-125 volts

INPUT IMPEDANCE: 10 MΩ and 25 μμf, below 10 millivolts; 10 MΩ and 8μμf above 10 millivolts

POWER SUPPLY: 105-125 volts; 50-420 cps, 75 watt. Provision for 210-250 volt operation

measures from

100 MICROVOLTS to 320 VOLTS

regardless of waveform

DIMENSIONS: *(Portable Model)* 14⅜" wide, 10⅛" high, 12⅜" deep—Relay Rack Model is available

WEIGHT: 21 lbs., approximately

Write for catalog for complete information

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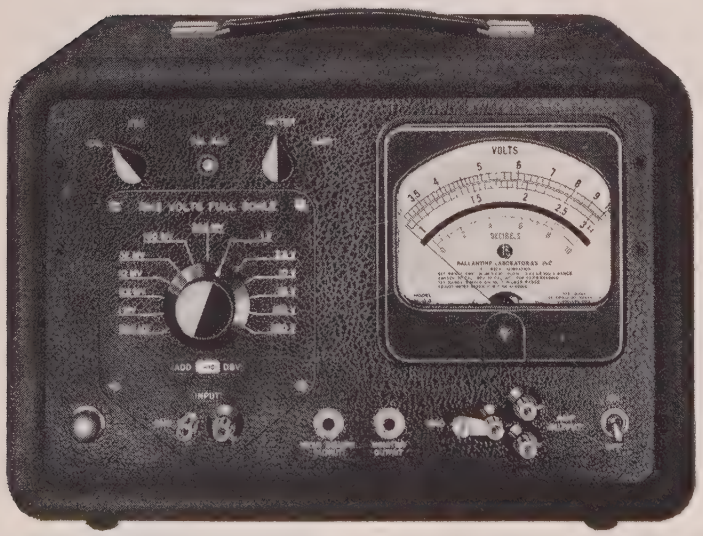

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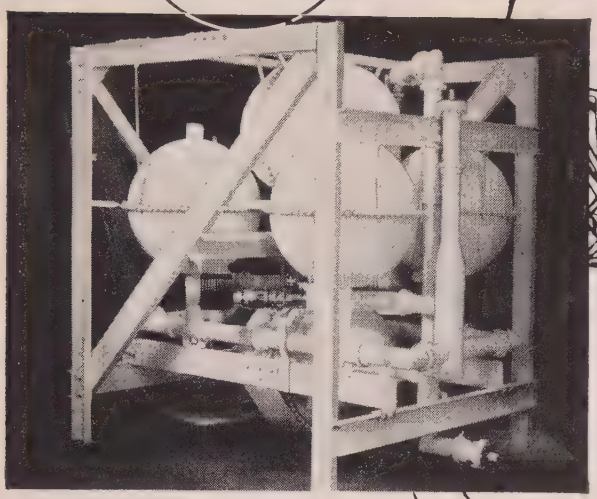

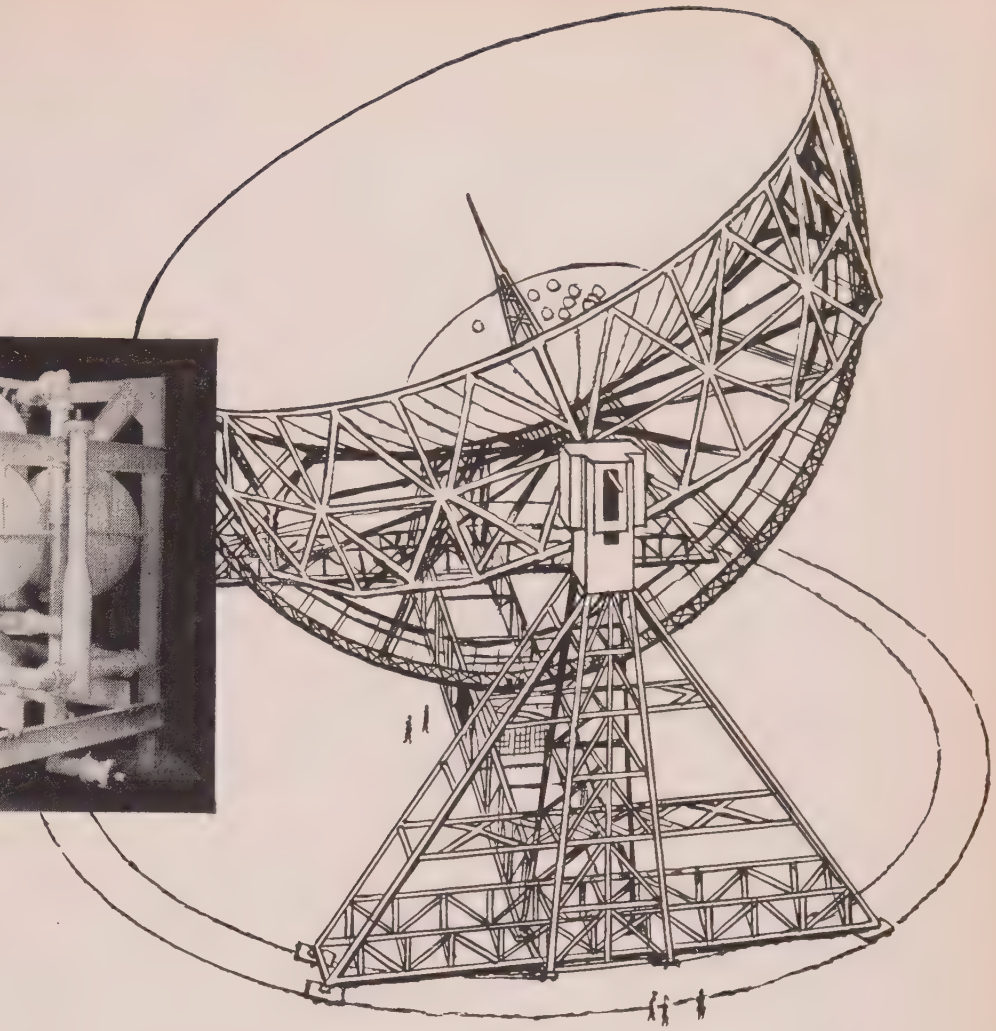

THIS UHF DIPLEXER NETWORK was built by **DIELECTRIC** Products Engineering Company, Inc., under contract from Space Technology Laboratories, Inc. It is now installed in the tower of the single giant antenna of "SpaN Net's" Jodrell Bank radio telescope at Manchester, England Its function is . . .

isolating simultaneous signals to and from deep-space payloads

Commanding and monitoring space vehicles is the responsibility of "SpaN Net" . . . a global network of radio telescopes set up and operated by Space Technology Laboratories for U.S.A.F.

To fulfill its mission, "SpaN Net" had to meet challenges far beyond the state of the communications art. Never before had powerful command signals and faint telemetered deep-space data been simultaneously handled by a single antenna. How could transmit and receive channels with a differential of more than 200 decibels be isolated? Conventional diplexer technology was not the answer.

Teamwork solved the isolation problem! "SpaN Net" assigned the task to Space Technology Laboratories, Inc., who added DIELECTRIC of Maine to its research, development and production team. Result? An ultra-high-frequency diplexer network that allows the Jodrell Bank radio telescope's single antenna to transmit 20 kw commands and receive 0.002 microvolt signals . . . simultaneously.

This contribution to SpaN Net's break-through is typical of product advances by DIELECTRIC If you have similar problems in communications . . . from design to delivery . . . it pays to contact DIELECTRIC. Capabilities, facilities and accomplishments are described in our brochure. Write for it today.

Other areas of **DIELECTRIC** capability in coaxial, waveguide and open wire techniques . . .

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dial DIELECTRIC for solutions to communications problems.

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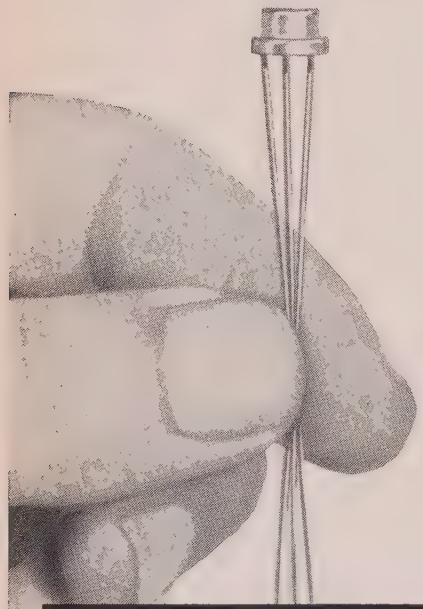
RAYMOND, MAINE

IERC TRANSISTOR HEAT DISSIPATOR

actual size

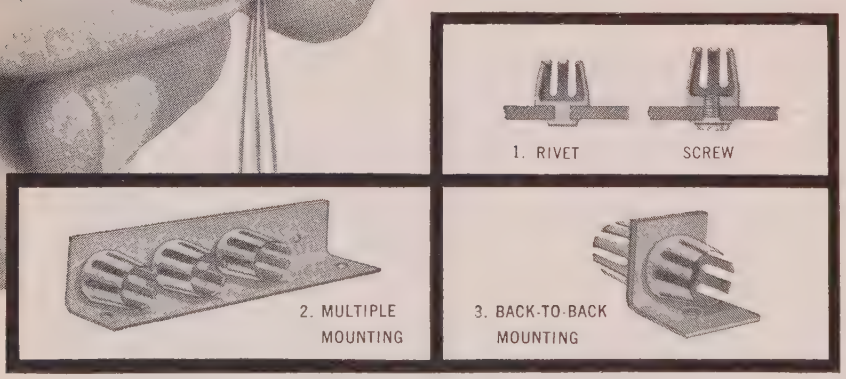
accepts .305 to .335 variations in TO-5 cases!



IERC Transistor Heat-dissipating Retainers readily accommodate diameter variations up to .030" found in TO-5, TO-9, TO-11, TO-39 transistor cases. This single IERC part saves you time and costs in specifying, stocking and application.

IERC's exclusive design features maximum thermal contact with transistor case for efficient transfer of heat to the dissipator and heat sink. Attaching methods suitable for printed circuit boards, chassis and heat sinks provide thermal benefits and retention in extreme shock and vibration environments.

Installation is a smooth, tension fit—eliminating the possibility of "snap-fit" impact injuries to the transistor!



Simplified installation for effective heat dissipation with IERC Transistor Heat Dissipators are illustrated: 1. Parts available in rivet or screw attaching types. 2. Single or multiple mounting on heat sink angle. 3. Back-to-back mounting.

Detailed information, performance graphs, etc. are available in latest IERC Technical Bulletin. Write for a copy today!

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Westinghouse Electric Corp., Semiconductor Department 906
Westline Products Division, Western Litho Co. 731
Weston Instruments Division, Daystrom, Inc. 305
White Sales Company, Inc. ..1143, 1144
Wood Electric Co., Inc. 931

Yewell Associates, Inc. 107, 1007

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ALL DIGITS CAN BE READ FROM ANY ANGLE

IN-LINE DIGITAL READOUT

featuring

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Series 10000

Over 1000 firms throughout the world in just a few years prove unprecedented acceptance of IEE digital readouts.

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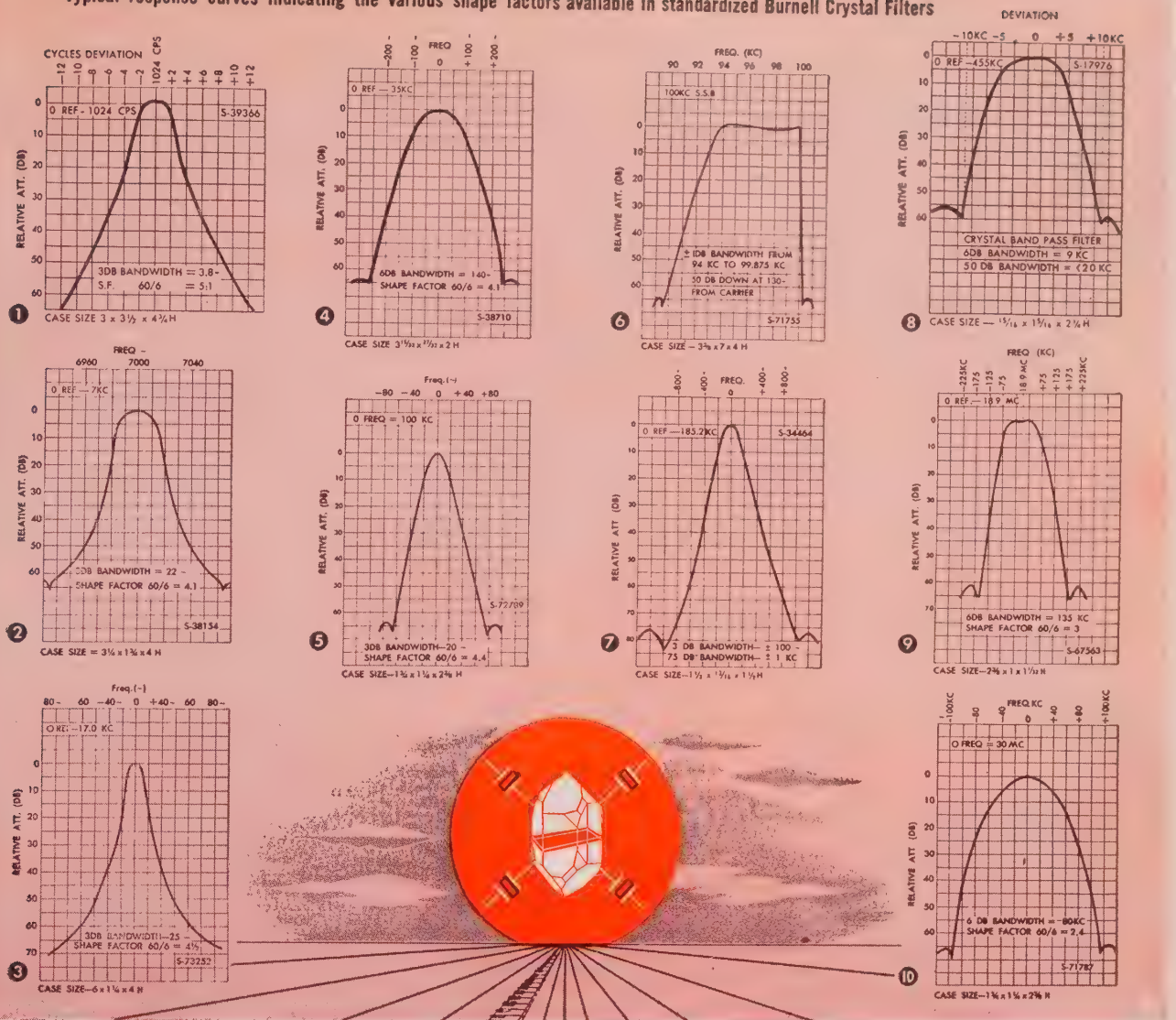
Typical response curves indicating the various shape factors available in standardized Burnell Crystal Filters


CYCLES DEVIATION
0 REF - 1024 CPS
S-39366
3DB BANDWIDTH = 3.8-
S.F. 60/6 = 5:1
CASE SIZE 3 x 3½ x 4¾H
FREQ
0 REF — 35KC
6DB BANDWIDTH = 140-
SHAPE FACTOR 60/6 = 4.1
S-38710
CASE SIZE 3 15/32 x 2 7/32 x 2 H
FREQ. (KC)
100KC S.S.B
± 1DB BANDWIDTH FROM
94 KC TO 99.875 KC
50 DB DOWN AT 130-
FROM CARRIER
S-71755
CASE SIZE — 3¾ x 7 x 4 H
DEVIATION
0 REF —455KC
S-17976
CRYSTAL BAND PASS FILTER
6DB BANDWIDTH = 9 KC
50 DB BANDWIDTH = <20 KC
CASE SIZE — 1 5/16 x 1 5/16 x 2¼ H
FREQ -
0 REF — 7KC
3DB BANDWIDTH = 22 -
SHAPE FACTOR 60/6 = 4.1
S-38134
CASE SIZE = 3¼ x 1⅜ x 4 H
Freq.(~)
0 FREQ = 100 KC
3DB BANDWIDTH—20 -
SHAPE FACTOR 60/6 = 4.4
S-72709
CASE SIZE—1⅜ x 1¼ x 2⅜ H
0 REF—185.2KC
S-34464
3 DB BANDWIDTH— ± 100 -
75 DB BANDWIDTH— ± 1 KC
CASE SIZE—1½ x 1 11/32 x 1½ H
FREQ (KC)
0 REF—18.9 MC
6DB BANDWIDTH = 135 KC
SHAPE FACTOR 60/6 = 3
S-67563
CASE SIZE—2⅜ x 1 x 1 11/32 H
Freq.(~)
O REF—17.0 KC
3DB BANDWIDTH—25 -
SHAPE FACTOR 60/6 = 4½
S-73252
CASE SIZE—6 x 1¼ x 4 H
FREQ.KC
O FREQ = 30 MC
6 DB BANDWIDTH = —80KC
SHAPE FACTOR 60/6 = 2.4
S-71787
CASE SIZE—1¾ x 1⅜ x 2⅜ H

Running the gamut in crystal filters

BURNELL CRYSTAL FILTERS NOW COVER FULLEST RANGE YET POSSIBLE

1 kc to 30 mcs

To its notable achievements in advancing the electronic arts, Burnell & Co. now adds another—the development of moderately priced high attenuation crystal filters covering the extraordinary range of 1 kc. to 30 mc. This represents a range many times broader than previously thought practicable. In addition, the Burnell Crystal Filter line now includes several types heretofore considered impossible.

More than 15 years research, development and experience are represented in the designs illustrated in the response curves shown. Burnell & Co. has taken crystal filters out of the luxury class in applying its experience to their design and manufacturing without incurring developmental and engineering costs.

Whether your crystal filter needs are for standard units or those engineered to center frequency, band width, selectivity and impedance level, call on our Crystal Filter Division for quick delivery. Send now for Crystal Filter Catalog, XT455.

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PELHAM, N. Y.
PELHAM 8-5000
TELETYPE PELHAM 3633

PACIFIC DIVISION
DEPT P-21
720 MISSION ST.
SOUTH PASADENA, CALIF.
MURRAY 2-2841
TELETYPE PASACAL 7578

Mfd. in Canada by Edo (Canada) Ltd., Cornwall, Ont. • WEllington 2-6774

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The KINTEL Model 501 4-digit, over-ranging digital voltmeter measures DC from ±0.0001 to ±1000.0 volts with 0.01% ±1 digit (of reading) accuracy. Ranging and polarity indication are automatic. The measured voltage is displayed in direct numerical form on an in-line, single-plane readout. Only an interconnecting cable is required to drive parallel entry digital printers. An extra fifth digit in the left decade indicates "0" or "1" to provide 100% over-ranging...ten times greater resolution at decade (1,10,100,1000) voltage points where other 4-digit voltmeters change ranges and lose a full digit of resolution. There's no better way to make fast, accurate DC voltage measurements. Price $2,995.

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ELECTRONICS, INC.

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Industrial Engineering Notes*

Engineering

Papers on the application of superconductive techniques to nformation processing devices in the United States and abroad presented at a Navy-sponsored symposium held last May in Washington have been released for sale to industry. Approximately 400 scientists and engineers currently engaged in research on cryogenic devices met at the symposium to exchange ideas and to learn the present status of the applications of superconductivity to computers, computing systems and information processing devices. The Office of Naval Research sponsored the symposium to focus attention on existing and future technological and theoretical problems that must be solved before practical devices can be developed. Some of the specific topics covered in the 57 illustrated papers are: closed cycle helium refrigeration, British research on superconductive switching devices, properties of thin film cryotrons, research on superconductive devices in Sweden, formation of thin polymer films by electron bombardment, and a computer program for stimulating cryotron circuits. The papers may be obtained by ordering PB 161763, "Symposium on Superconductive Techniques for Computing Systems," Office of Naval Research, May, 1960, from the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C. The volume contains 418 pages, price $4.50. The Office of Technical Services has also published a 28-page Army report reviewing major programs in electronic microminiaturization underway at educational, industrial and government laboratories. Order PB 161674 "Status of Electronic Microminiaturization." 75 cents. . . . **The reports of electronics R&D reviewed below were published recently by the Office of Technical Services as part of its continuing program of collection, reproduction and dissemination of technical information generated by the defense program.** The reports may be ordered from OTS, U. S. Department of Commerce, Washington 25, D. C., at the prices shown. "Research in Radiation Damage in Semiconductors," J. W. Harrity and others, General Dynamics Corporation for U. S. Air Force. February, 1960. 157 pages. (Order PB 161673. $3.) A diode resistant to nuclear radiation damage was constructed in this research program as the result of extensive studies of nuclear irradiation effects on semiconducting materials. The diode uses p-type germanium with resistivity of about 0.20 ohm-cm as the base material. Principal object of the study was to determine the mechanisms by which radiation degrades semiconductor devices. Conductivity, Hall effect, and carrier lifetime of germanium and silicon were studied as a function of total irradiation between 78 degrees and 300 degrees K. The researchers state that one of the most significant accomplishments of the program was the development of the instrumentation and techniques of measurement used with the General Atomic electron linear accelerator. Results acquired by their experiments demonstrate that performing detailed measurements while the sample is being irradiated can be accomplished effectively and with minimum use of expensive accelerator and reactor time. "High-Temperature Semiconductor Devices," edited by F. D. Rosi and P. Rappaport, Radio Corporation of America for U. S. Air Force. October, 1959. 124 pages. (Order PB 161541. $2.75.) This work for the Air Force is to some extent a continuation and extension of research in the field of compound semiconductors and devices undertaken at RCA Laboratories in 1953 through which gallium arsenide and indium phosphide emerged as the most promising semiconductors for high-temperature applications. This report contains a detailed description of the scientific results obtained in research programs on these materials, basic physical measurements, and devices. The section on materials reports progress in the purification, crystal growth, and evaluation of the compounds GaAs and InP, with major emphasis on GaAs. The section on basic physical measurements contains reports on determination of impurities in GaAs, diffusion in GaAs and InP, and minority carrier lifetime in GaAs. The report on devices describes accomplishments in the fabrication of diodes, rectifiers and transistors of both GaAs and InP. "Searching the Literature for Transducer Information: Part 1—A Guide to the Literature," J. Pearlstein, Diamond Ordnance Fuze Laboratories. October, 1959. 59 pages. (Order PB 161320. $1.75.) This technical literature survey on transducers in ordnance equipment consists of first principles in classifying related phenomena and materials and developing criteria for documentation. It is a revision of an earlier study and prerequisite to a new Army Ordnance program designed to develop better sensory devices.

Goverment and Legislature

Total funds for the performance of basic research in the United States reached a level of $1 billion in the year 1959–1960, according to estimates released recently by the National Science Foundation. The billion-dollar estimate is based on the NSF statistical series, "Funds for the Perform-

(Continued on page 44A)

* The data on which these Notes are based were selected by permission from *Weekly Report*, issues of August 29 and September 6 and 12, 1960, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

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is experienced... in Airborne Navigation Systems

The hummingbird is well-equipped by nature to home and hover.

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LFE's AN/APN-78 was the first, fully-automatic, self-contained Doppler Navigator specifically designed for helicopters. Continuing developments have led to a full line of systems for ASW and AEW aircraft, capable of accurate, long range navigation *and* accurate hovering indication.

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NAME YOUR OSCILLOGRAPHIC RECORDING NEEDS

INPUT CHARACTERISTICS

FREQUENCY RESPONSE RANGE ... RECORDING METHOD

PACKAGING ... NUMBER OF CHANNELS

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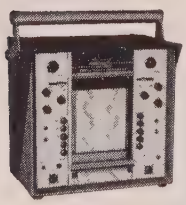
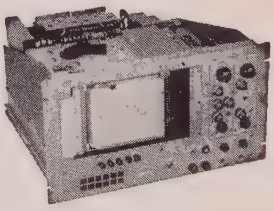
"BUILDING-BLOCK" FLEXIBILITY CAN MEET THEM

1-channel, 21 lb. portables — Model 299 for general purpose DC recording, Model 301 for AC strain gage recording.

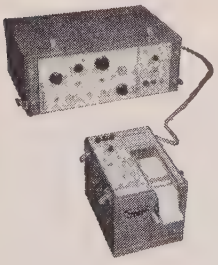


Model 320 2-channel, 4-speed portable . . . 50 mm wide channels.

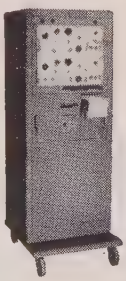


Model 297 2-channel recorder, rack mount in 10½" space or portable housing.

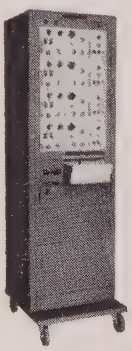
1- to 8-channel "150's", DC to 100 cps.



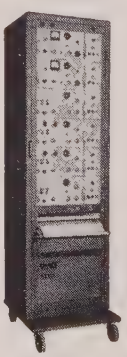
MODEL 151



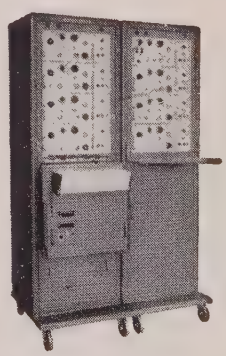
MODEL 152



MODEL 154



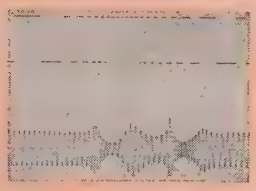
MODEL 156

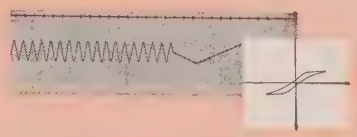


MODEL 158

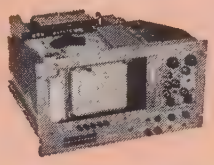
Interchangeable, plug-in preamps for "850" systems include Carrier, Phase-Sensitive Demodulator, DC Coupling, Low Level . . . for "350" systems, all "850" types plus 400 Cycle Frequency Deviation, Logarithmic types . . . for "150" systems, all "350" and "850" types plus AC-DC, AC Wattmeter, RMS Volt/Ammeter, Frequency Meter, Stabilized DC types, and Triplexer 3-channel electronic switch.

Single-chassis, 6- and 8-channel amplifiers for "950" systems include high and low gain general purpose types . . . for "650" systems and optical oscillographs of other manufacturers, a 6- or 8-channel medium gain amplifier.





From DC to 150 cps with most Sanborn systems which produce inkless, heated stylus recordings in true rectangular coordinates . . . *up to* 5000 *cps* with new optical system using ultraviolet-sensitive paper. *Optical X-Y Recorder* has 2500"/sec. writing speed, immediate readout.



Sanborn "building block" units give you extremely wide packaging flexibility: "150" amplifiers and recorders can be separately housed or combined in mobile cabinets . . . "350" preamps may be used alone or in a system . . . "850" and "950" systems can have up to 16 channels in a single cabinet . . . optical 5000 cycle recorder can be housed with X-Y recorder, etc. *All* Sanborn multi-channel recorder assemblies and 350, 650, 850 & 950 amplifier modules fit standard 19" racks.

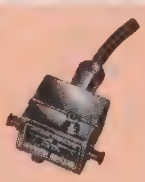
Up to 8 brilliant, long-persistence traces presented simultaneously on the new 17" Sanborn Model 769 scope. Circuit includes individual gating amplifiers for each Y axis.

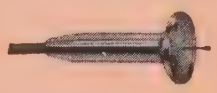
SANBORN COMPANY

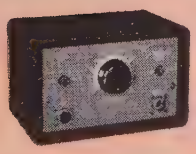
INDUSTRIAL DIVISION

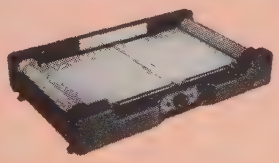
175 Wyman Street, Waltham 54, Mass.

Measure linear velocity and displacement with Sanborn LVsyn, Linearsyn and "probe-style" transducers. Strokes from ±0.005" to 10", high sensitivities, 0.5% linearity; special types for high temperatures, pressures . . . single-ended or differential pressure measurements at 40 μv/0.1mmHg/volt sensitivities, with 267, 268 Series pressure transducers.

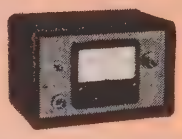




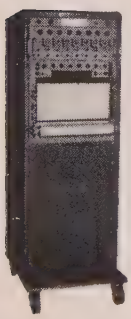




Complete Sanborn equipment for your oscillographic recording needs includes a variable speed *Chart Viewer* . . . *Servo Monitor Phase Shifter* . . . "150" Series *Wide Band Driver Amplifier* . . . a *Monitor Meter* . . . 4- to 32- channel *Event Recorders* with *Transistor Driver Amplifiers.*

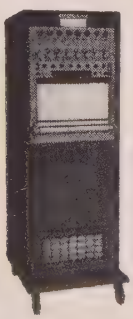


See how Sanborn oscillographic recording building-block flexibility can provide the most suitable equipment for *your* needs. Ask your Sanborn Sales-Engineering Representative — offices in principal cities throughout the United States, Canada and foreign countries.



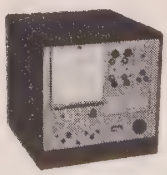
Economical 8-channel "950" high gain, general purpose system.

Model 958-16 Complete 16-channel system with amplification . . . cabinet or rack mounting.

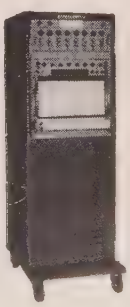


6- or 8-channel "850's", min. panel space.

Model 650 1- to 24-channel optical system, DC to 5000 cps.



Model 670A High speed (100 cps) optical X-Y recorder.



Simplified 6- or 8-channel system, 5 volts full scale.

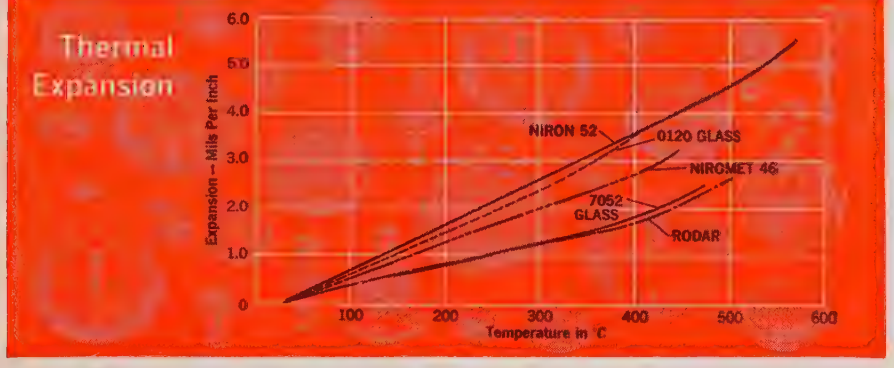
6- or 8-channel "350's", DC to 150 cps.

specialized alloys for glass hermetic seals

RODAR® NIRON® 52 NIROMET® 46

Specified industry-wide for

PERMANENTLY-BONDED VACUUM-TIGHT SEALS!



RODAR®

NOMINAL ANALYSIS: 29% Nickel, 17% Cobalt, 0.3% Manganese, Balance—Iron

Rodar matches the expansivity of thermal shock resistant glasses, such as Corning 7052 and 7040. Rodar produces a permanent vacuum-tight seal with simple oxidation procedure, and resists attack by mercury. Available in bar, rod, wire, and strip to customers' specifications.

Temperature Range		Average Thermal Expansion *cm/cm/°Cx10-7	
30° To 200°C.		43.3 To 53.0	
30	300	44.1	51.7
30	400	45.4	50.8
30	450	50.3	53.7
30	500	57.1	62.1

COEFFICIENT OF LINEAR EXPANSION
*As determined from cooling curves, after annealing in hydrogen for one hour at 900° C. and for 15 minutes at 1100° C.

NIRON® 52

NOMINAL ANALYSIS: 51% Nickel, Balance—Iron

This Wilbur B. Driver nickel iron alloy contains 51% nickel. Niron 52 sealing alloy is exceptionally well adapted, and widely employed, for making seals with 0120 glass.

NIROMET® 46

NOMINAL ANALYSIS: 46% Nickel, Balance—Iron

A 46% nickel-balance iron alloy with expansion properties between Niron 52 and Rodar. It is used extensively as terminal bands for vitreous enameled resistors.

Call or write for Sealing Alloy Bulletin

WILBUR B. DRIVER CO.
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IN CANADA: Canadian Wilbur B. Driver Company, Ltd.
50 Ronson Drive, Rexdale (Toronto)

Precision Electrical, Electronic, Mechanical and Chemical Alloys for All Requirements

Industrial Engineering Notes

(Continued from page 40A)

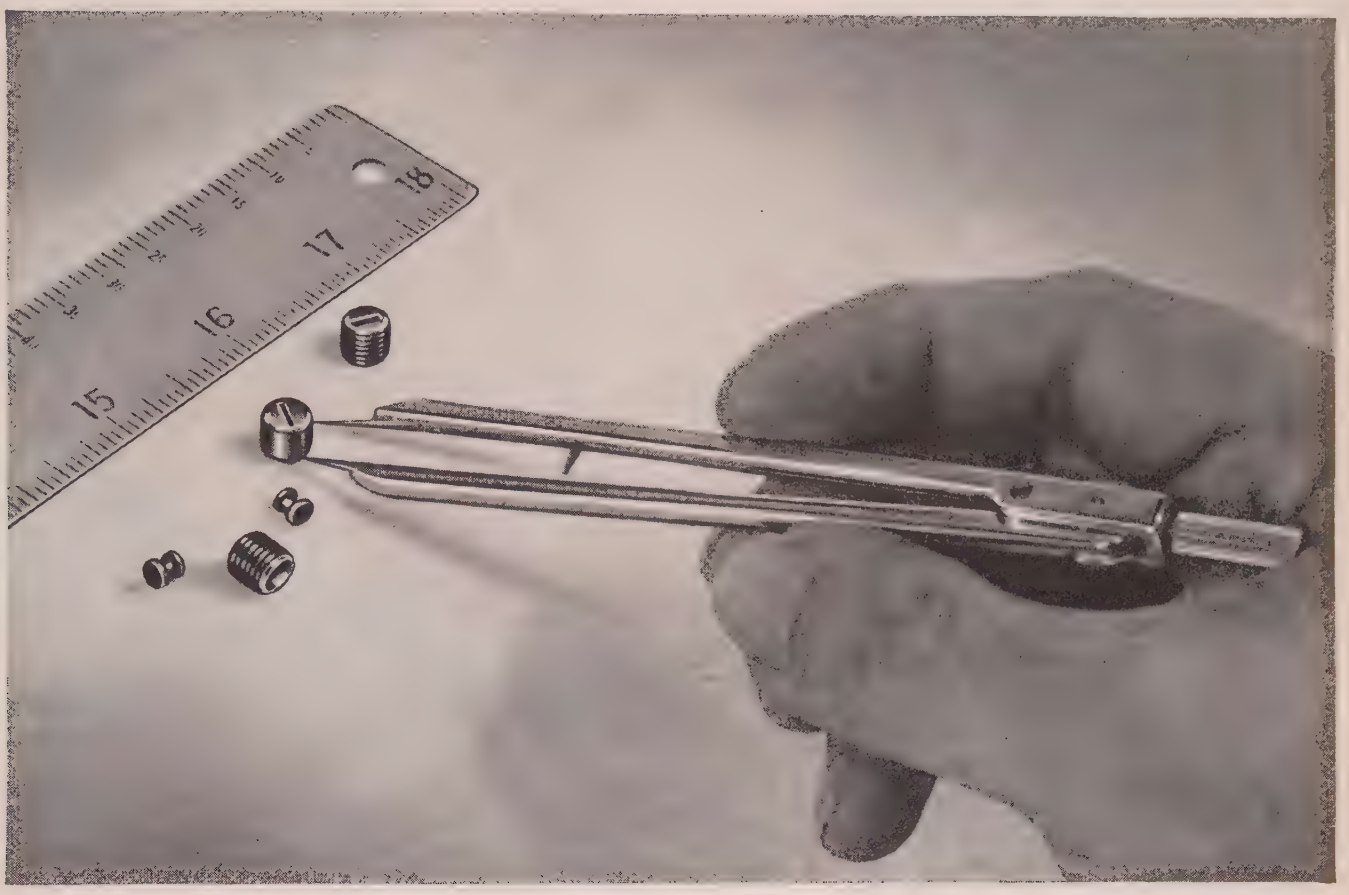
ance of Basic Research in the United States, 1953–85, Reviews of Data on Research & Development, No. 22." Basic research funds rose from about $430 million in 1953–54 to more than $830 million in 1957–58, an increase of 93 per cent. . . . **The Federal Aviation Agency will spend some $34.2 million during fiscal year 1961 on projects involving electronics, according to an analysis of the FAA budget released by the EIA Marketing Data Department.** The projects to be undertaken with research and development funds include studies and engineering services and some hardware. For installation and modification of air navigation facilities, $195 million was granted in new obligational authority. Some $95.6 million is earmarked for radars, communications and other electronic areas. . . . **The Office of Civil and Defense Mobilization, acting on advice of the military services, has closed the door on plans which the Federal Communications Commission hoped would solve one the most politically hot problems it has on frequency allocations.** On August 19 OCDM Director Leo A. Hoegh informed the FCC that both of the Commission's plans for exchanging commercial and government channels to obtain more VHF television channels had been rejected as too risky to national security "in today's international climate." As a result of this apparent final decision, the FCC is back where it was several years ago when the Senate Interstate and Foreign Commerce Committee began demanding that it come up with a solution to the television allocation problem resulting from too few VHF frequencies and the unpopularity of the many UHF channels it has for disposal. The OCDM decision consequently is expected to force the Commisssion either to reconsider various plans for mixing UHF and VHF allocations or otherwise to implement the use of UHF. In this connection, the Commission is now preparing for UHF experiments in New York City under a Congressional grant during the next two years. . . . **Almost 9 of every 10 U. S. households have television sets, according to results of a sample survey conducted earlier this year by the Bureau of the Census.** The survey, conducted at the request and expense of the Advertising Research Foundation, shows that the increase in households with television sets continued through 1959 and early 1960. Eighty-eight per cent of all households had television sets as compared with 86 per cent in May, 1959, 83 per cent in January, 1958, and 12 per cent in April, 1950. The popularity of television is indicated not only by the spread of television to more households,

(Continued on page 48A)

Use your IRE DIRECTORY! It's valuable!

ANOTHER G-C SPACE SAVER!

Threaded Cup Core and Bobbin for Miniaturized IF Transformer and Coil Applications



Now you can design miniaturized IF transformers and coils for AM-FM radio and television applications, and maintain high "Q" and effective permeability factors.

This new G-C threaded cup core and ferrite bobbin assembly has closely held mechanical tolerances for precision screw adjustment, and allows close coupling factor between the coil and core.

For RF coil applications, G-C offers a complete line of threaded cores operating in frequency ranges up to 65 mcs. G-C threaded cores are available from stock in Q-1 and Q-2 material; pitch sizes from ¼–28, 10–32 and 8–32; lengths from ¼" to 1⅛" with hex, square or screw-driver holes.

Write for additional information on the miniaturized threaded cup core F1266 and bobbin F1270 and data on G-C stock cores. Please address inquiries to Section P.

DIVISION OF INDIANA GENERAL CORPORATION

GENERAL CERAMICS

KEASBEY, NEW JERSEY, U.S.A.

TECHNICAL CERAMICS, FERRITE AND MEMORY PRODUCTS

One of a series

Catching the Drift of Gyro Bearings

The fantastic accuracies needed by inertial guidance systems for space flight depend on the suppression of gyro drift, the tendency of a gyro to precess from minutely occurring internal torques. Particularly puzzling has been the problem of "jogs," or sudden axial shifts, within gyro spin-axis bearings. Shifts of but one ten-millionth of an inch can cause serious steering error.

Specialists at the GM Research Laboratories have found that the real key to drift lies in the thickness and distribution patterns of bearing lubricating films. Only a tenth of a milligram of oil — equivalent in volume to less than two-thousandths of a drop of water — is required in a gyro bearing, but even this amount unevenly distributed may cause jogs.

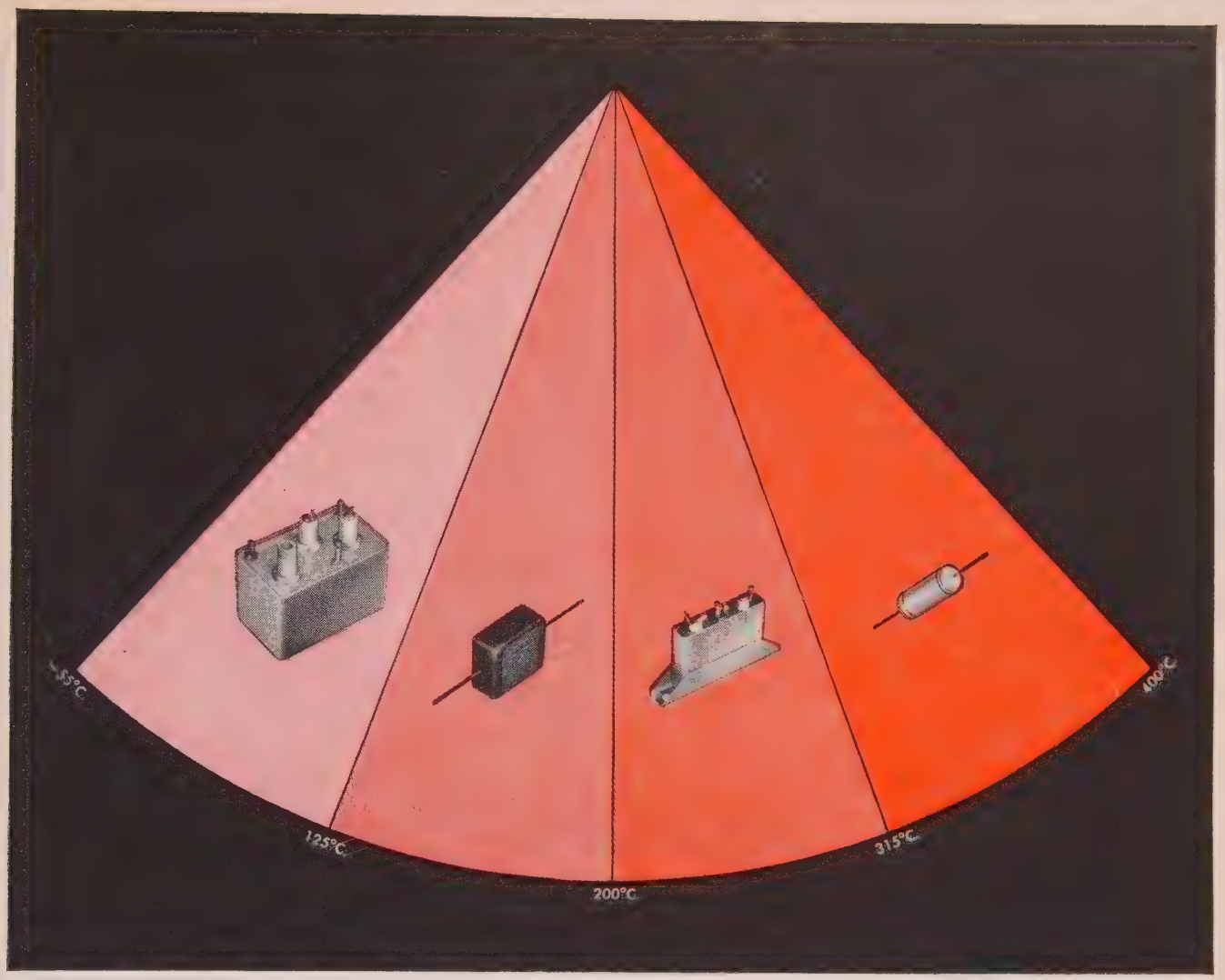
Conducting unique studies of single bearings apart from rotor assemblies, GM Researchers use a hydrostatic spindle and special instrumentation to take film-thickness measurements they compare with hydrodynamic theory. Jogs, due to excess oil supply, have been analyzed in relation to surface oil transfer and separator feed control, ball spin orientation, displacement, and differential heating and ball wander.

This experimental and analytical approach is achieving progress toward jog-free, stably distributed, and suitably thick oil films required in high-precision bearings. It is a further example of the critical and advanced research General Motors carries out in seeking "more and better things for more people."

The fluoresced streaks show the disturbed "wake" of the lubricating film during bearing operation. The active part of the film, too thin to fluoresce visibly, averages ten-millionths of an inch in thickness.

General Motors Research Laboratories

Warren, Michigan



BENDIX CAPACITORS COVER A FULL TEMPERATURE SPECTRUM

ALL FEATURE THESE IMPORTANT ADVANTAGES:

Environmental resistance **Wide voltage range** **High I. R.** **Radiation resistant**
No voltage derating **Solid impregnants** **Wound mica papers** **Exceptional stability**

Under 125°C.—Specials
• Size and weight reductions at high voltages • Drift—.25% capacitance change typical from −55°C. to +125°C. • High I. R.—1500 megohm X microfarads typical at 125°C. • Solid impregnants—no liquid leakage.

125°C. to 200°C.—Available soon
• .001 to 6.0 mfd., 200 V to 3 KV, specials to 10 KV. • Molded and metal housed; tubular and rectangular • Size and weight reduction—over plastic film and stacked mica types, particularly at high voltages • Drift—1% capacitance change typical from −55°C. to +200°C. • High I. R.—50 megohm X microfarads typical at 200°C. • Proved in 4 years' usage.

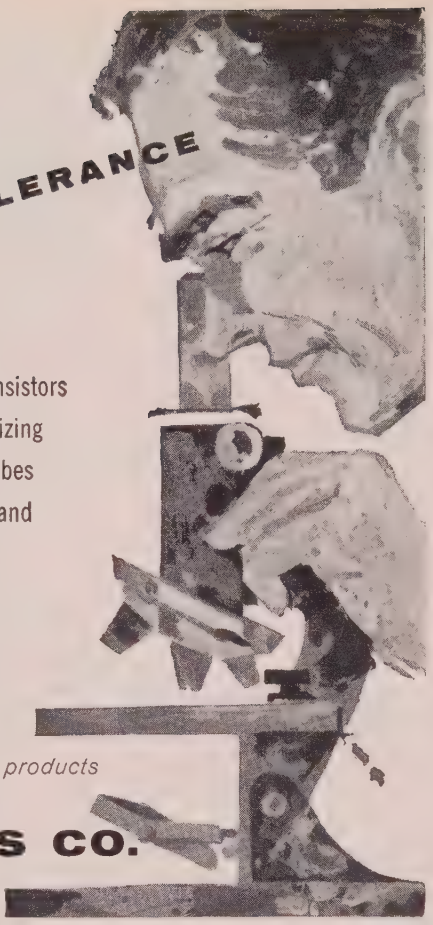
200°C. to 315°C.—In production
• .05 to 4.0 uf, 600 V and up • Drift—3% capacitance change typical from −55°C. to +315°C. • High I. R.—10 megohm X microfarads typical at 315°C. • Nothing smaller at 315°C.

315°C. to 400°C.—In development
• .001 to 6.0 uf, 150 V and 600 V • Drift—5% capacitance change typical from −55°C. to +426°C. • High I. R.—1 megohm X microfarad typical at 400°C. • Prototype availability • Only inorganic materials used.

For full details, write:

Scintilla Division
Sidney, New York

THE Bendix CORPORATION



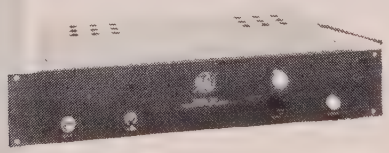
ALL IN MICRON TOLERANCE

Evaporation masks for Mesa transistors
Micro Mesh Sieves for particle sizing
Aperture masks for color T.V. tubes
Electroformed mesh for storage and image tubes
Etched glass and metal reticles
Etched intricate metal parts
Flexible printed wiring
Miniaturized printed circuits

For precisely controlled dependable products call or write . . .

BUCKBEE MEARS CO.
ST. PAUL 1, MINNESOTA

Wide Band Amplifier Model 530



GENERAL DESCRIPTION

The Model 530 Wide Band Amplifier has been designed to fill a need for an amplifier used principally for voltage amplification of CW or pulsed signals.

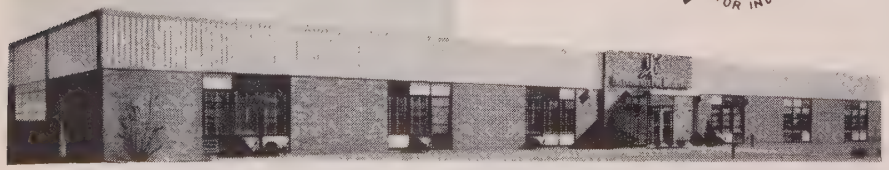
This model has many applications in the laboratory and also for television distribution systems. It may be used to increase the output from signal sources within its frequency range, and its bandwidth of 300 mc's makes it ideal for amplifying millimicrosecond pulses.

PRICE $330.

SPECIFICATIONS

Bandpass	10 KC to 300 MC
Voltage Gain	18 db
Input Impedance	135 ohms
Output Impedance	150 ohms
Max. Output Power (into Matched Load)	.08W
Max Output Voltage (into Matched Load)	3.5 Vrms
Max. Peak Pluse Output	7 V pos. or neg.
Rise Time	Less than $2x10^9$ sec
Dimensions	19" front panel—16" wide, 9" deep 3½ high. Power supply included.
Gain Control	Provided on front panel
Tube Complement	Two cascaded stages of eight 6AK5

INSTRUMENTS FOR INDUSTRY, Inc.
101 New South Road, Hicksville, L. I., N.Y.



Graduate engineers with two or more years of circuit application in the fields of electronics or physics are invited to meet with Mr. John Hicks in an informal interview or send complete resume to: Dir. Personnel, IFI, 101 New South Road, Hicksville, New York.

Industrial Engineering Notes

(Continued from page 44A)

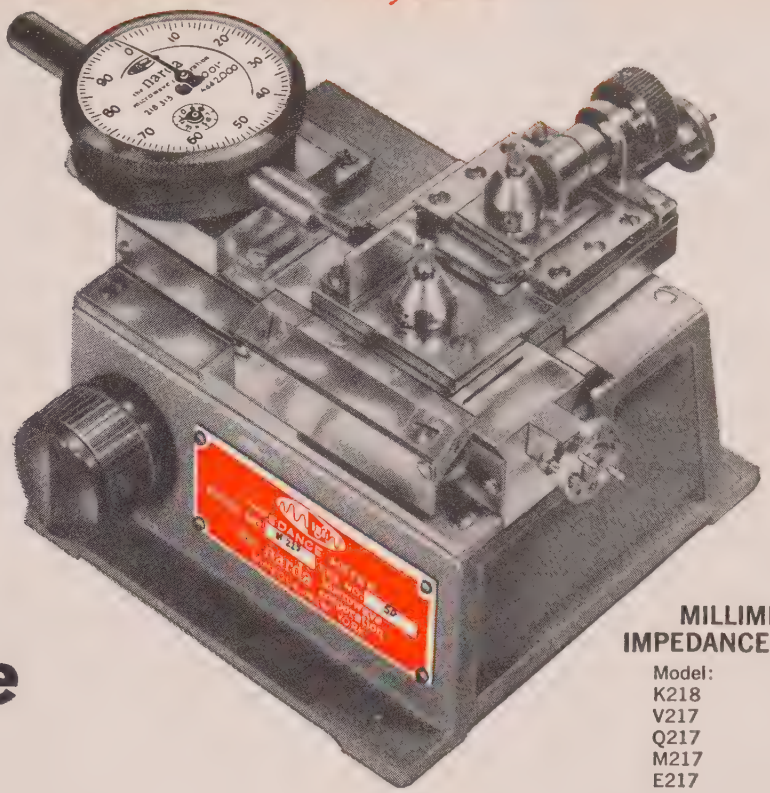
but also by the growing number of households with more than one set. Eleven per cent of all households had two or more sets, compared with 8 per cent in May, 1959, and 7 per cent in January, 1958. Households with one set constituted 77 per cent of the total, while only 12 per cent were without television. As in the earlier surveys of the series, television sets were found more common inside standard metropolitan statistical areas (SMSA's) than outside these areas. Among SMSA households, 91 per cent had one or more sets, while the percentage for areas outside SMSA's was 82 per cent. By geographical region the northeast was highest with 92 per cent of its households reporting television sets. The north central and the west were next, with 89 per cent and 87 per cent, respectively. In the south, 82 per cent of households had television. Television among rural farm households continued to increase: 76 per cent had television sets compared with 73 per cent in May, 1959, and 68 per cent in January, 1958. Similar increases were reported for rural nonfarm areas and places not in urbanized areas. Among households with a telephone available, 94 per cent had television sets, while 66 per cent of households without a telephone available had one or more sets. . . . **An industry committee aiding the Federal Communications Commission in its UHF-TV testing program is reported to have agreed that New York City's Empire State Building is the only feasible location for the main transmitting antenna.** The General Industry Advisory Committee (GIAC), meeting in Washington, reportedly approved the technical recommendations of the Transmitter Industry Advisory Committee for location of the main antenna. The GIAC also is reported to have recommended the use of a 360° antenna. If cost or technical difficulties preclude its use, a single 180° or two 180° antennas were recommended. EIA President L. Berkley Davis and Frederick R. Lack, Director of the Engineering Department, are among industry representatives on the General Industry Advisory Committee. The group advises the FCC on policy and procedural matters involved in the UHF testing program. . . . **Monthly statistics for transistors showed a sharp drop in both sales and revenue at the factory.** Year-to-date totals, however, remained far ahead of those for the same period in 1959. The figures are shown in the following table.

TRANSISTOR SALES (UNITS)

	Factory Sales (Units)
July	7,070,884
June	10,392,412
May	9,046,237
April	9,891,236
March	12,021,506
February	9,527,662
January	9,606,630
Year-to-date '60	67,556,567
Year-to-date '59	42,128,291

Working from 18,000 to 90,000 mc?

Don't settle for less than narda precision Impedance Meters!



MILLIMETER IMPEDANCE METERS

Model:	Price:
K218	$ 500.
V217	600.
Q217	1100.
M217	995.
E217	1650.

Just because you're working in millimeter bands is no reason to "make do" with lesser specs than you get from L to KU Bands. It's true that it's more difficult to maintain top specs at 90,000 mc than at X Band, for example, but in most cases Narda does it! And especially, in our slotted lines!

Precision instruments in every way, Narda millimeter Impedance Meters give you these important benefits: Smooth, stable carriage travel, absolutely parallel to the transmission line axis, with complete freedom from backlash. Imperceptible slope (0.01 max.), thanks to precision milling of tellurium copper main transmission line, followed by a silver plate of molecular thickness and a rhodium flash.

Residual VSWR is under 1.01 for K and V Bands; 1.02 for Q, M and E. Position of the built-in probe is measured with a precision dial indicator having a least count of 0.001". And as an added advantage, the probe is usable with any waveguide detector.

All in all, an exceptionally fine instrument — and yet, typical of all Narda millimeter products. For complete information on the entire line, write for a free catalog. Address: Dept. PIRE-6.

Rely on narda for these other precision millimeter products:

Direct Reading VARIABLE ATTENUATORS
HARMONIC GENERATORS
High-Directivity DIRECTIONAL COUPLERS
FREQUENCY METERS
SLIDING TERMINATIONS
FIXED TERMINATIONS
TUNABLE SHORTS
CRYSTAL DETECTORS
STANDARD GAIN HORNS

the narda microwave corporation

118-160 HERRICKS ROAD, MINEOLA, L. I., N. Y. · PIONEER 6-4650

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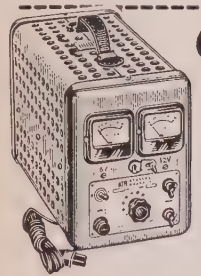
ATR PLUG-IN TYPE PORTABLE INVERTERS*

A.C. HOUSEHOLD ELECTRICITY Anywhere . . . in your own car!

Operates Standard A.C.
- Record Players
- Dictating Machines
- Small Radios
- Electric Shavers
- Heating Pads, etc.

In your own car or boat!

MODELS
6-RMF (6 volts) 60 to 80 watts. Shipping weight 12 lbs. DEALER NET PRICE $33.00
12T-RME (12 volts) 90 to 125 watts. Shipping weight 12 lbs. DEALER NET PRICE $33.00
*Additional Models Available



ATR "A" Battery ELIMINATOR

For Demonstrating and Testing Auto Radios—
TRANSISTOR or VIBRATOR OPERATED!

Designed for testing D.C. Electrical Apparatus on Regular A.C. Lines—Equipped with Full-Wave Dry Disc-Type Rectifier, assuring noiseless, interference-free operation and extreme long life and reliability.

MAY ALSO BE USED AS A BATTERY CHARGER
MODEL 610C-ELIF . . . 6 volts at 10 amps. or 12 volts at 6 amps. Shipping weight 22 lbs.
DEALER NET PRICE $49.95
MODEL 620C-ELIT . . . 6 volts at 20 amps. or 12 volts at 10 amps. Shipping weight 33 lbs.
DEALER NET PRICE $66.95

AUTO-RADIO ATR VIBRATORS

By every test ATR Auto-Radio Vibrators are best! . . . and feature Ceramic Stack Spacers, Instant Starting, Large Oversized Tungsten Contacts, Perforated Reed, plus Highest Precision Construction and Workmanship and Quiet Operation!

There is an ATR VIBRATOR for every make of car!

Ask your distributor for ATR's Low Priced type 1400, 6 volt 4-prong Vibrator; and 1843, 12 volt 3-prong; or 1840, 12 volt 4-prong Vibrator. THE WORLD'S FINEST!

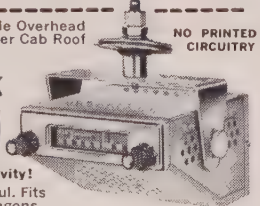
There is a trim plate kit for YOUR CAR!

ATR CUSTOMIZED KARADIO
Vibrator-Operated with Tone Control

ATR KARADIO . . . is ideal for small import cars or compact American cars! Unit is completely self-contained—extremely compact! Powerful 8-tube performance provides remarkable freedom from engine, static, and road noises. The ATR Customized Karadio comes complete with speaker and ready to install. Can be mounted in-dash or under-dash—wherever space permits! No polarity problem. Neutral Gray-Tan, baked enamel finish. Overall size, 7" deep, 4" high, and 6½" wide. Shipping weight, radio set, 7 lbs.
Model K-1279—12 for 12V Dealer Net Price . . $33.57
Model K-1279— 6 for 6V Dealer Net Price . . $33.57

Airplane Style Overhead Mounting under Cab Roof

NO PRINTED CIRCUITRY



ATR TRUCK KARADIO

Excellent Tone, Volume, and Sensitivity!

Compact, yet powerful. Fits all trucks, station wagons, most cars and boats. Just drill a ⅝ inch hole in roof and suspend the one-piece unit (aerial, chassis and speaker) in minutes. Watertight mounting assembly holds antenna upright. Yoke-type bracket lets you tilt radio to any angle.

Extra-sensitive radio has 6 tubes (2 double-purpose), over-size Alnico 5 PM speaker for full, rich tone. Big, easy-to-read illuminated dial. Fingertip tuning control. Volume and tone controls. 33-in. stainless steel antenna. Neutral gray-tan enameled metal cabinet, 7 x 6½ x 4 in. high over-all. Shipping weight 10½ lbs.
Model TR-1279—12A for 12V Dealer Net Price $41.96
Model TR-1279— 6A for 6V Dealer Net Price $41.96

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Quality Products Since 1931
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IRE People

Pierre J. Tapernoux (S'47–A'50–SM '58) has been appointed system design manager in the Systems Management Department of Stromberg-Carlson's Electronics Division.

P. J. TAPERNOUX

In this capacity, he is responsible for the design of a complex $27-million electronic reconnaissance system being built by General Dynamics Corporation's Stromberg-Carlson Division for the U. S. Air Force.

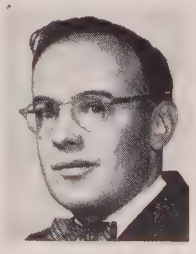
He has been with Stromberg-Carlson for 12 years, having joined the company in 1948 as an electronic engineer in the Research Division. Subsequently, he served as project engineer in charge of the development of miniaturized airborne voice communication equipment, as project manager for the development of a large digital control system for controlling a supersonic wind tunnel nozzle structure, as manager of the Radio Communication Laboratory, and, for the past two years, as head of the Systems Analysis Section of the Electronic Systems Department.

During World War II, he served in the U. S. Army Signal Corps as a radio operator, later attended officer training school and became a First Lieutenant in the Army Security Agency, where he had charge of circuit development and testing of speech privacy devices.

Mr. Tapernoux received the electrical engineering degree from the College of the City of New York in 1948. He holds several patents, most of them classified.

❖

Dr. Harold F. Webster (SM'58) of the General Electric Research Laboratory, Schenectady, N. Y., will be an official U. S. delegate to the General Assembly of the International Scientific Radio Union to be held in London, September 5–15. As an official delegate, he will present a review of United States work on energy conversion and act as official reporter for the conference discussions on energy conversion. While in Europe, he will also visit laboratories in England, Norway, Sweden, Germany and Switzerland.



H. F. WEBSTER

Dr. Webster joined the Research Laboratory in 1951, specializing in physical electronics and microwave technology. He was a co-recipient of the 1958 W. R. G. Baker Award for the outstanding paper published in the TRANSACTIONS for the professional groups. His recent work has been primarily in the field of direct conversion of heat to electricity by thermionic devices.

❖

James J. Burns (M'57) has been named to the new position of District Sales Representative for the Los Angeles office of Motorola Semiconductor Products Inc., according to a recent announcement.

J. J. BURNS

He was formerly an electronics development engineer at Goodyear Aircraft Corporation in Litchfield Park, Ariz. He was also a member of the U. S. Army Signal Corps during the Korean War.

In his new position, he will be responsible for customer engineering services for Motorola's western district.

A native of Pennsylvania, Mr. Burns received the B.S.(EE) degree from the Milwaukee School of Engineering, Milwaukee, Wis.

❖

George Caryotakis (S'50–A'51–M'57) has joined Eitel-McCullough, Inc., San Carlos, Calif., manufacturer of electron-power tubes, as senior project engineer, Power Klystron Laboratory. He is responsible for high-power broadband klystron development in the company's Power Klystron Division at San Bruno, Calif. Dr. Caryotakis was formerly senior engineer, Tube Research Department, at Varian Associates.

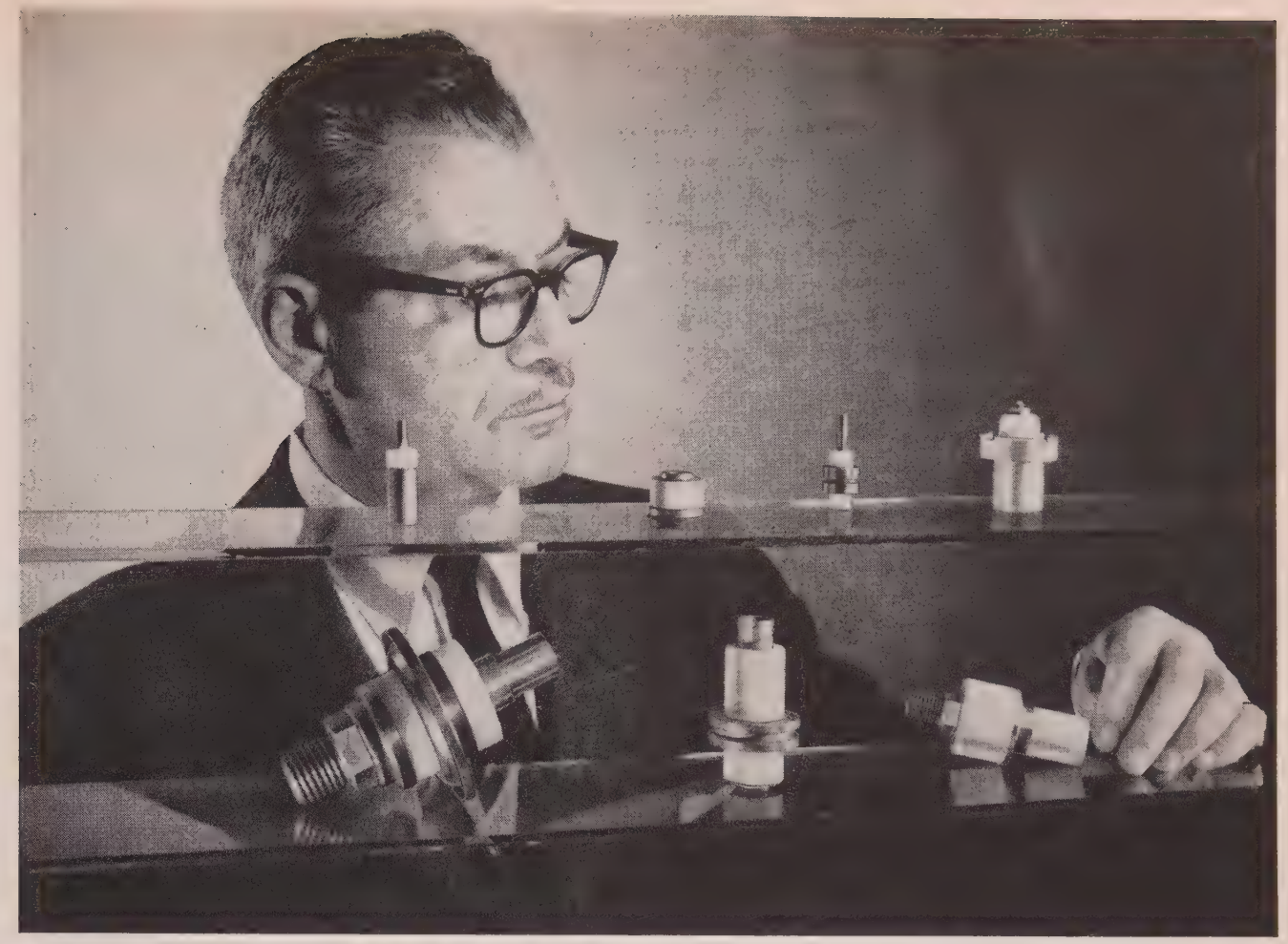
G. CARYOTAKIS

After receiving the B.S. degree in electrical engineering from Syracuse University in 1951, Dr. Caryotakis obtained the M.S. and Ph.D. degrees in the same field from Stanford University in 1952 and 1955, respectively. The following year he worked on periodic focusing of traveling-wave tubes at Stanford Electronics Research Laboratories as a research associate.

Dr. Caryotakis has several patents pending on klystron design and has presented a number of papers on this subject. He is a member of Sigma Xi and Tau Beta Pi.

❖

(Continued on page 54A)



Engineering hints from Carborundum

Use KOVAR® Alloy to solve problems in sealing to ceramics

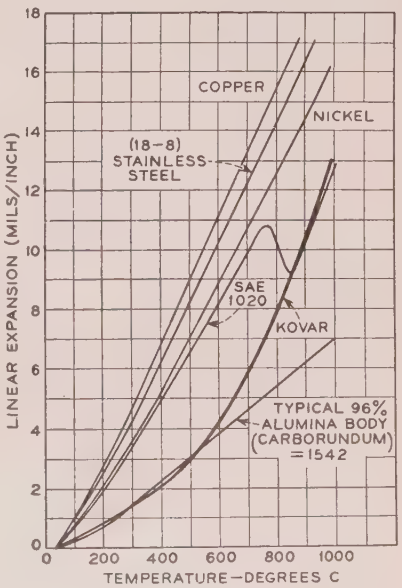
KOVAR, the original 29% nickel, 17% cobalt, 54% iron alloy, was developed for sealing to low expansion glass, but is now being used extensively for making pressure and vacuum tight seals with metallized ceramics of the low expansion type.

The curves at right show the expansion of KOVAR compared with a representative high alumina ceramic body. The expansivity match up to 500 C is very close, and the difference in expansion at higher temperatures is closer than with most common metals and alloys.

The fact that KOVAR is slightly higher in expansion at elevated temperatures is an actual advantage when the ceramic is on the inside of the unit since the resulting joint is placed in compression. The degree of compression is slight compared with that resulting from the use of a metal of higher expansion.

While a considerable difference in expansivity can sometimes be tolerated with the metal on the outside of thick sections of ceramic, this is not the case when the ceramic section is thin. Closer compatibility of expansivity, such as is obtained with KOVAR, is also required when the metal is on the inside of the ceramic or for sandwich or end type seals where both tensional and shear stresses must be kept to a minimum.

KOVAR alloy is stocked in a variety of sizes of rod, wire, tubing, sheet, cups and eyelets. Your inquiries are invited for prices, technical information and recommendations on specific problems. Write Dept. P-110 Latrobe Plant, Refractories Division, The Carborundum Company, Latrobe, Pa.



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CARBORUNDUM®

EXCELLENT form-factor and operating versatility make these rugged magnetrons ideal for many small-package applications including CW or pulsed radar beacons, test equipment oscillators, airborne navigation, proximity detection, surveillance, and transponder type operations.

Light, dependable, and with proven capabilities, these tubes operate at 500 to 600 peak volts and 150 ma peak pulsed current, permitting low-cost modulator components for all applications. They give a nominal power output of 1 watt CW and 15 watts peak.

Engineering programs in progress at Microwave Associates are directed towards development of this tube as a voltage-tunable magnetron within the same form-factor. Your inquiries are welcomed on these and other magnetrons.

A copy of our new 72 page Magnetron Catalog is available upon written request on your company letterhead.

MICROWAVE ASSOCIATES, INC.
BURLINGTON, MASSACHUSETTS

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IRE People

(Continued from page 50A)

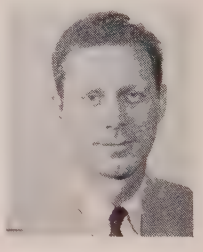
Herbert W. Bass (M'55) has recently joined F & F Associates, Electronic Manufacturers Representative, as a third partner.

H. W. Bass

He has held sales managerial positions with two manufacturers representatives and a well-known manufacturer of AC instruments in the East since 1957, and will cover the key areas of New York City and Westchester. Prior to 1957, he spent six years with the Sperry Gyroscope Co., holding the title of Engineering Group Leader at the time he left them to enter the sales field.

❖

Robert T. Cavanagh (S'44–A'46–M'55–SM'56)has been appointed General Manager of Philips Electronic Instruments, Division of Philips Electronic & Pharmaceutical Industries Corp., Mount Vernon, N. Y., it has been announced.


R. T. Cavanagh

He joined the Philips organization early this year and held the position of Assistant General Manager until he received his new assignment. Born in Winnipeg, Canada, he received the B.A.Sc. degree in Electrical Engineering from the University of Toronto, Toronto, Canda, where he majored in Communications and Electronics. He moved to the U. S. in 1947 and became a citizen in 1957.

From 1947 to 1960, he was with Allen B. DuMont Laboratories, Inc., where he was General Manager of Military Electronics Operations (1959–1960), Director of Research and Development (1956–1958), Director of Circuit Research Division (1953–1956), Chief Engineer of Receiver Division (1951–1953), and Assistant to Director of Research (1947–1951).

His past experience has involved Radar and Communications, Data and Display, Nuclear Instrumentation, Television, Reconnaissance, Missile Support Equipment, and Product Design. His responsibility as General Manager of Military Electronics Operations for DuMont involved the management of four company divisions.

Credited with several patents, Mr. Cavanagh holds membership in the American Rocket Society, American Astronautical Society, Armed Forces Communications & Electronics Association, and the Association of Professional Engineers of the Province of Ontario.

❖

(Continued on page 58A)

Will The Mission Succeed?

Ask . . .

DATS

(Dynamic Accuracy Test System)

Flight-line checkout by DATS (Dynamic Accuracy Test System) tells the interceptor commander whether his aircraft and weapon control systems are completely ready for a successful mission. As a result of field evaluation tests, showing the effectiveness of DATS in improving weapon control performance, RCA has been awarded an Air Force production contract. Developed by RCA's Airborne Systems Division, Defense Electronic Products, Camden, New Jersey, DATS is a new approach to the evaluation of system readiness.

It makes certain that only aircraft with properly operating weapon control systems are sent on missions. Based on a building-block design employing the highest reliability factors, a mechanical programming device and self-test capability, DATS utilizes a series of synthesized attack runs typical of mission conditions. DATS could be made applicable to many interceptor types of aircraft.

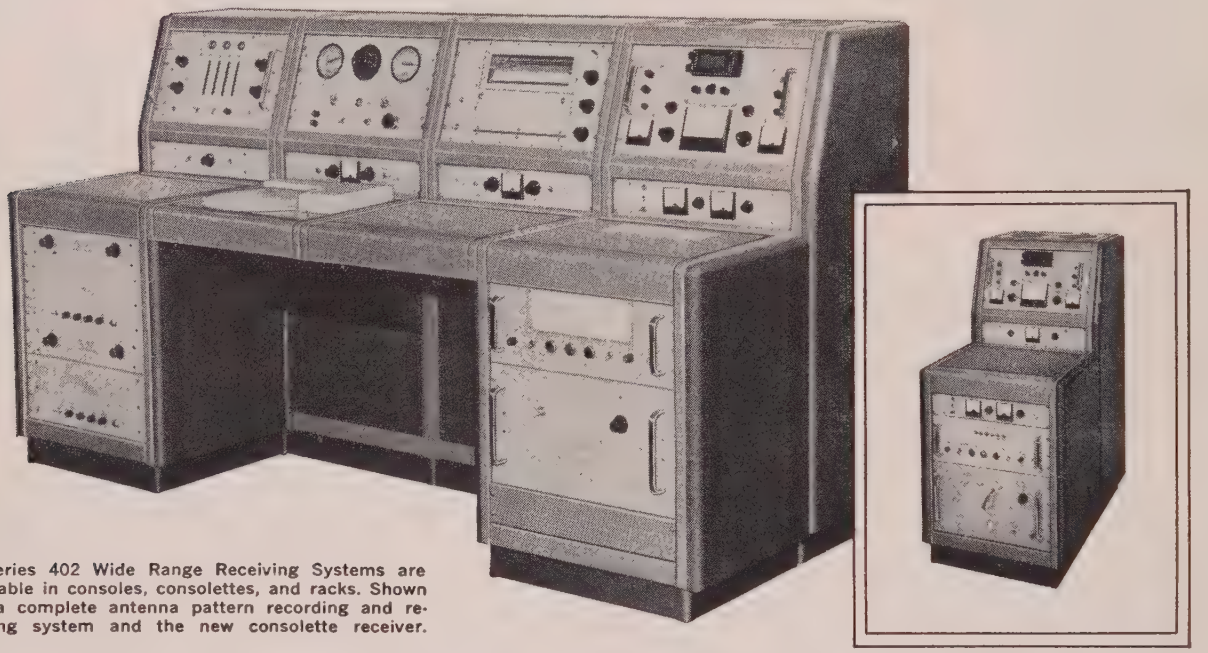
RCA®

The Most Trusted Name in Electronics

RADIO CORPORATION OF AMERICA

MAJORS and MINORS

...A Message to the Antenna Designer



● Series 402 Wide Range Receiving Systems are available in consoles, consolettes, and racks. Shown are a complete antenna pattern recording and receiving system and the new consolette receiver.

A crowded spectrum plus high power radar and communication systems critically compound the problems of the antenna design engineer.

More than ever, the complete pattern including **all** the major and minor lobes of every radiating element must be graphed for sound engineering evaluation.

S-A Receiver Gets the Whole Signal

Scientific-Atlanta Series 402 Wide Range Receiving Systems are specifically designed for antenna pattern measurements. Unique in design, these receivers combine maximum sensitivity and linearity from 30 mc to above 100 kmc. They are also useful as multipurpose laboratory instruments for microwave testing, monitoring, and measuring applications.

Only from S-A, 1 db Linearity over Full 60 db Dynamic Range

A recent development, S-A's P-4 modification adds 20 db to the normal 40 db dynamic range. The modification takes advantage of the gain vs AGC voltage characteristics of the Series 402. Existing receivers can be modified at the factory.

New Modification Z Broadens Use

Modification Z adds a precision IF attenuator and VTVM to the Series 402. Now RF and microwave signal level, gain, and isolation measurements can be made with fewer components and instruments. *For instance,* an X band 80 db attenuator can be calibrated to within ±0.5 db with a 1 mw signal source, a flap attenuator, a mixer, and an S-A Series 402Z Receiver. Antenna gain can be measured by direct comparison with a standard gain antenna. Signal levels can be compared against a reference standard.

Other Features

One coaxial cable from antenna to receiver eliminates costly lossy waveguides and rotary joints. Antenna can be located up to 75 feet away with negligible loss in sensitivity ☆ One receiving system covers 30 mc to above 100 kmc without plug-ins ☆ Reception of cw signals from simple sources eliminates need for precise modulation ☆ High sensitivity means low source power and long ranges ☆ High selectivity reduces interference and cross talk between adjacent test ranges ☆ Positive AFC action over full dynamic range provides pattern recording in deep nulls.

PRICES

Series 402, 2 to above 100 kmc	$7500
Series 402A, 2 to above 100 kmc with AGC	8000
Series 402B, 30 mc to above 100 kmc	8500
Series 402C, 30 mc to above 100 kmc with AGC	9000
Modification P-4	500
Modification Z	1000

NEW DATA FOLDER READY

For complete information ask for our new data folder from your nearby S-A engineering representative or write directly to Dept. 86.

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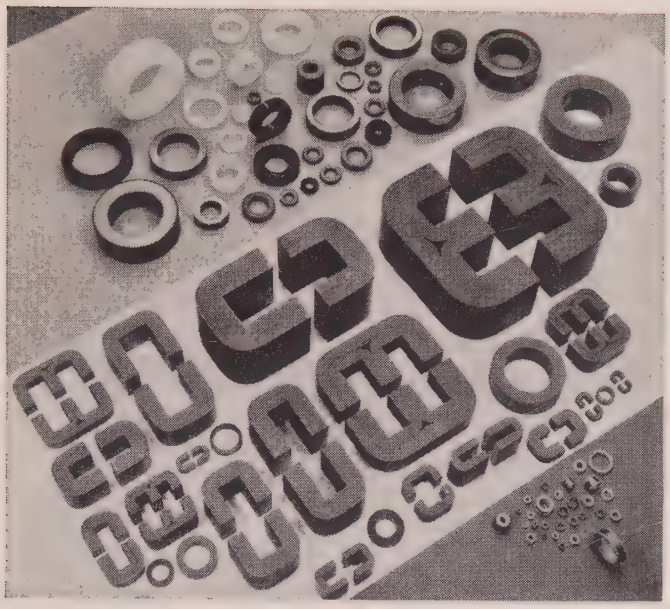
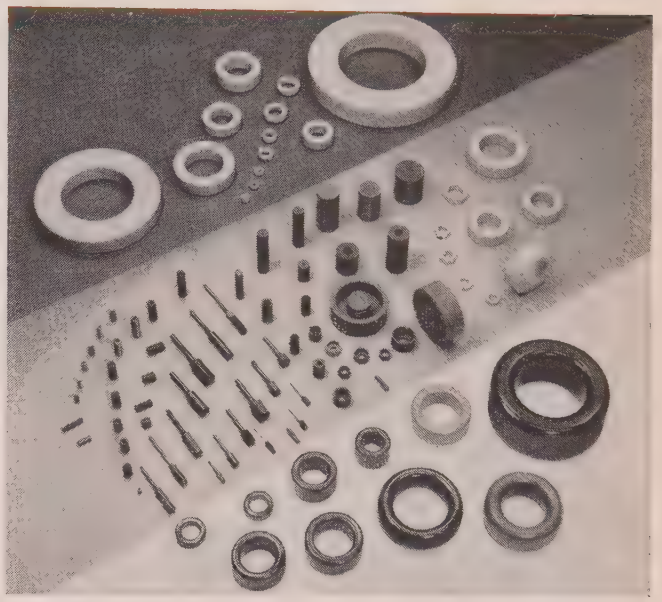
3

Specify "ARNOLD" for your MAGNETIC CORE requirements



Top to bottom: Tape wound cores, Silectron C, E and O cores, and bobbin cores.



Top to bottom: Mo-Permalloy powder cores, iron powder cores, and Sendust cores.

SILECTRON C-CORES, E-CORES and TOROIDS Arnold C and E cores are made from precision-rolled Silectron strip in 1, 2, 4 and 12 mil thicknesses.

They are supplied in a wide variety of shapes, and in sizes from a fraction of an ounce to several hundred pounds. In addition to standard transformer applications, they may also be supplied for special applications such as saturable reactors, instrument transformers and pulse transformers.

Over 1,000 stock cores are listed in the Arnold Silectron catalog. A wide selection of preferred sizes are carried in stock for immediate shipment. For complete data on C and E cores and Silectron toroids, *write for Bulletin SC-107A.*

TAPE WOUND CORES of High Permeability Materials Arnold tape wound cores are available made of Deltamax, 4-79 Mo-Permalloy, Supermalloy, Mumetal, 4750 Electrical Metal, Silectron, or the new rectangular-loop material, Supermendur. All except Supermendur cores are available in standard tape thicknesses of 1, 2 and 4 mils; also in special tape thicknesses of ½ mil, 12 mil or other, as required or feasible. Supermendur is presently available in 4 mil cores.

Toroidal cores are made in 30 standard sizes with protective nylon or aluminum cases. Special sizes of toroidal cores are produced to individual requirements. *Write for Bulletin TC-101A. (TC-113A for Supermendur Cores.)*

BOBBIN CORES Arnold bobbin cores are available in a wide range of sizes, tape thicknesses, widths and number of wraps to suit the ultimate use of the core in electronic computer assemblies. Magnetic materials usually employed are Deltamax and Square Permalloy in standard thicknesses of 1, ½, ¼ and ⅛ mil. Bobbins are supplied in ceramic or stainless steel. *Write for Bulletin TC-108A.*

MO-PERMALLOY POWDER CORES Available in a wide range of sizes, from .260" OD to 5.218" OD. They are given various types of enamel and varnish finishes, some of which permit winding with heavy Formex insulated wire without supplementary insulation over the core.

These powder cores are supplied in four standard permeabilities: 125, 60, 26 and 14 Mu. They provide constant permeability over a wide range of flux density, and in many cases may be furnished stabilized to provide essentially constant permeability over a specific temperature range. Large warehouse stocks of preferred sizes are carried for immediate shipment. *Write for Bulletin PC-104C.*

IRON POWDER CORES A wide selection of cores is available, from simple cylinders to special cores of complicated design. The line includes all standard types of threaded cores, cup, sleeve, slug and cylindrical insert cores: for use in antenna and RF coils, oscillator coils, IF coils, perm tuning, FM coils, television coils, noise filter coils, induction heating and bombarder coils, and other low frequency applications. Preferred sizes are carried in warehouse stock for quick shipment. A standard series of iron powder toroids is also manufactured, conforming to the standard sizes proposed by the Metal Powder Industries. *Write for Bulletin PC-109.*

SENDUST POWDER CORES Available in a wide selection of sizes, ranging from .800" OD to 3.346" OD, and in permeabilities of 10, 13, 25, 30, 50 and 80, although not all sizes are available in all permeabilities. They possess magnetic properties generally superior to iron powder cores, but inferior to Mo-Permalloy powder cores in the audio and carrier frequency range. *Write for Bulletin SDC-110.*

SPECIAL MATERIALS

2V PERMENDUR . . . a ferromagnetic alloy of cobalt, vanadium and iron that possesses high flux density saturation properties. Its magnetostrictive properties are useful in many transducer applications. *Write for Bulletin EM-23.*

VIBRALLOY . . . a ferromagnetic alloy of nickel, molybdenum and iron whose temperature coefficient of elastic modulus is controllable over a wide range. It has high ferromagnetic permeability, and a rather high coefficient of magnetostriction. Used in applications where a zero or controlled thermo-elastic coefficient is desired.

BARIUM TITANATE . . . A piezoelectric ceramic widely used in accelerometers, phono pickups, microphones, ultrasonic grinding and cleaning devices and underwater signaling devices. For more information, *write for Bulletin CM-116.*

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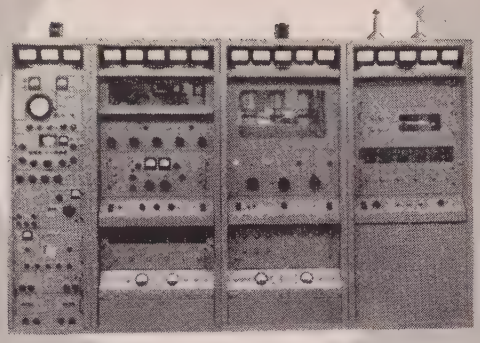
- ALDC
- Filtered air cooling
- Semi-pressurized cabinet
- Safety interlocks throughout



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- DSB*
- CW
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- FS
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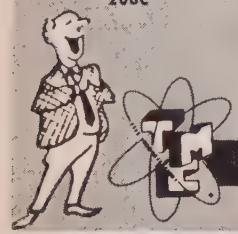
The TMC transmitter, Model GPT-40K, is a completely self contained unit including all power supplies and ventilating equipment. All components are housed in four modular assemblies occupying only 40 square feet of floor space.

The transmitter includes equipment for immediate monitoring and testing of all vital operating points.

To provide for complete flexibility in operation, 1 Kw or 10 Kw outputs are readily available for reduced power or emergency operation.

The unit is available with or without synthesizer control.



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A frequency standard so accurate that it measures time with a rate of change of less than one second in sixty years!

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A JK-SULZER FS1100T currently serves as the basic quartz crystal Frequency Standard employed by WWVL (Bureau of Standards, Boulder, Colorado) to provide a 20kc reference signal having a stability of 1 x 10^{10}/Day.

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SPECIALISTS IN FREQUENCY MANAGEMENT for space exploration programming, high speed navigation, and spectrum conservation in the growing communications field.

IRE People

(Continued from page 54A)

John W. Coltman (SM'49–F'55), formerly manager of the electronics and nuclear physics department of Westinghouse research laboratories, has been named an associate director of the laboratories.

In his new assignment, he will head the company's central research program on advanced equipment concepts, including research in new methods of power generation and propulsion, special types of electronic tubes, solid state devices and electromechanical devices and control systems.

He was born in Cleveland, Ohio, where he received the B.S. degree in physics from the Case Institute of Technology in 1937. He earned the M.S. and Ph.D. degrees at the University of Illinois, Urbana, and joined the Westinghouse research laboratories in 1941 when he was awarded a Westinghouse Research Fellowship.

During World War II, he was engaged in radar research, particularly in the development of the magnetron tube that was essential to radar transmission. In 1946, he was named to head a group to undertake the development of electronic techniques for increasing the brightness of X-ray images.

One of his major contributions was the invention, in collaboration with Dr. F. Marshall, of the scintillation counter, a device which superseded the Geiger counter as a detector of nuclear radiation.

Since 1949, as head of the electronics and nuclear physics department, Dr. Coltman has been guiding research in infrared and optical imaging, underwater acoustics, microwaves, electron physics and nuclear physics. He is a fellow of the American Physical Society.

❖

The appointment of **Herman Fialkov** (A'47–M'55) as president of General Instrument's Semiconductor Division has been recently announced. He was a founder of General Transistor and headed that company as president since 1954 and until its recent merger into General Instrument.

H. FIALKOV

Among other appointments resulting from the merger are: Vice President-Marketing of the Semiconductor Division, **Kenneth C. Moritz** (M'53–SM'58). He joined General Instrument earlier this year; he previously was National Sales Manager for the Raytheon Company's Semiconductor Division.

Dr. Bernard Jacobs (M'56) was named Vice President for Research and Development. He was formerly Vice President for research of General Transistor, and will now be in over-all charge of the Semi-

(Continued on page 62A)

Another New Achievement from El-Menco

A New Smaller Size

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Mica Capacitor

El-Menco's SUB-MINIATURE DM-10 Mica Capacitor...

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Smaller than a 1-carat diamond!

Sets New Standard in Miniature Reliability!

This sub-miniature DM-10 Mica Capacitor retains the same superior electrical characteristics of silvered mica capacitors as found in much larger sizes. It assures a high order of performance in extreme miniaturization applications — missiles, printed circuits and all compact electronic equipment. Parallel leads provide greater versatility. Tough phenolic casings protect against physical damage and penetration of moisture.

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Working Voltage	Capacity Range
100 WVDC	1 MMF thru 360 MMF
300 WVDC	1 MMF thru 300 MMF
500 WVDC	1 MMF thru 250 MMF

Operating Temperature: *up to 150° C.*

Characteristics: *C, D, E and F, depending on capacitance value*

Leads: *#26 AWG (.0159") Copperweld wire*

EL-MENCO'S DM-10 MEETS ALL THE ELECTRICAL REQUIREMENTS OF MILITARY SPEC. #MIL-C-5B AND EIA SPECIFICATION RS-153

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DM-19 . . . up to 5400 mmf at 300 VDCW, up to 4000 mmf at 500 VDCW.

WRITE FOR SAMPLES OF EL-MENCO DM-10 CAPACITORS and brochures describing El-Menco's complete line of capacitors.

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28 Fields of Special Interest-

The 28 Professional Groups are listed below, together with a brief definition of each, the name of

Aeronautical and Navigational Electronics

Annual fee: $2.

The application of electronics to operation and traffic control of aircraft and to navigation of all craft.

Mr. William P. McNally, Chairman, 35 Laurel St., Floral Park, L.I., N.Y.

36 Transactions, *6, & *9, and Vol. 2, No. 1-3; Vol. 3, No. 2; Vol. 4, No. 1, 2, 3; Vol. 5, No. 2, 3, 4; Vol. 6, No. 1, 3, 4; Vol. 7, No. 1, 2.

Antennas and Propagation

Annual fee: $4.

Technical advances in antennas and wave propagation theory and the utilization of techniques or products of this field.

Prof. Edward C. Jordan, Chairman, Electrical Engineering Dept., University of Illinois, Urbana, Ill.

32 Transactions, *Vol. AP-2, No. 2; AP-4, No. 4; AP-5, No. 1-4; AP-6, No. 1, 2, 3, 4; AP-7, No. 1, 2, 3, 4; AP-8, No. 1, 2, 3, 4, 5.

Audio

Annual fee: $2.

Technology of communication at audio frequencies and of the audio portion of radio frequency systems, including acoustic terminations, recording and reproduction.

Mr. H-S. Knowles, Chairman, Knowles Electronics, 9400 Belmont Ave., Franklin Park, Ill.

54 Transactions, *Vol. AU-1, No. 6; *Vol. AU-2, No. 4; Vol. AU-3, No. 1, 3, 5; Vol. AU-4, No. 1, 5-6; Vol. AU-5, No. 1, 2, 3, 4, 5, 6; AU-6, No. 1, 2, 3, 4, 5, 6; AU-7, No. 1, 2, 3, 4, 5, 6; AU-8, No. 1, 2, 3, 4.

Automatic Control

Annual fee: $3.

The theory and application of automatic control techniques including feedback control systems.

Mr. John M. Salzer, Chairman, 909 Berkeley St., Santa Monica, Calif.

10 Transactions, PGAC-3-4-5-6, AC-4, No. 1, 2, 3; AC-5, No. 1, 2, 3.

Bio-Medical Electronics

Annual fee: $3.

The use of electronic theory and techniques in problems of medicine and biology.

Dr. Herman P. Schwan, Chairman, University of Pennsylvania, School of Elec. Engrg., Philadelphia 4, Pa.

17 Transactions, 8, 9, 11, 12; ME-6, No. 1, 2, 3, 4; ME-7, No. 2, 3.

Broadcast & Television Receivers

Annual fee: $4.

The design and manufacture of broadcast and television receivers and components and activities related thereto.

Mr. Robert R. Thalner, Chairman, Sylvania Home Electronics, Batavia, N.Y.

26 Transactions, *7, 8; BTR-1, No. 1-3, BTR-2, No. 1-2-3; BTR-3, No 1-2; BTR-4, No. 2, 3-4; BTR-5, No. 1, 2, 3; BTR-6, No. 1, 2.

Broadcasting

Annual fee: $2.

Broadcast transmission systems engineering, including the design and utilization of broadcast equipment.

Mr. George E. Hagerty, Chairman, Westinghouse, 122 E. 42nd St., New York 17, N.Y.

17 Transactions, No. 2, 6, 7, 10, 11, 12, 13, 14; BC-6, No. 1, 2, 3.

Circuit Theory

Annual fee: $3.

Design and theory of operation of circuits for use in radio and electronic equipment.

Mr. Sidney Darlington, Chairman, Bell Tel. Labs., Murray Hill, N.J.

27 Transactions, CT-3, No. 2; CT-4, No. 3-4; CT-5, No. 1, 2, 3, 4, CT-6, No. 1, 2, 3, 4; CT-7, No. 1, 2.

Communications Systems

Annual fee: $2.

Radio and wire telephone, telegraph and facsimile in marine, aeronautical, radio-relay, coaxial cable and fixed station services.

Capt. C. L. Engleman, Chairman, Engleman & Co., Inc., 2480 16th St., N.W., Washington 9, D.C.

18 Transactions, CS-5, No. 2, 3; CS-6, No. 1, 2; CS-7, No. 1, 3, 4; CS-8, No. 1, 2.

Component Parts

Annual fee: $3.

The characteristics, limitation, applications, development, performance and reliability of component parts.

Mr. Floyd E. Wenger, Chairman, Headquarters ARDC, Andrews AFB, Washington 25, D.C.

20 Transactions, CP-4, No. 1, 2, 3-4; CP-5, No. 1, 2, 3, 4; CP-6, No. 1, 2, 3, 4; CP-7, No. 1, 2.

Education

Annual fee: $3.

To foster improved relations between the electronic and affiliated industries and schools, colleges, and universities.

Dr. John G. Truxal, Chairman, Dept. of EE, PIB, Brooklyn, N.Y.

11 Transactions, Vol. E-1, No. 3, 4; E-2, No. 1, 2, 3, 4; E-3, No. 1, 2, 3.

Electron Devices

Annual fee: $3.

Electron devices, including particularly electron tubes and solid state devices.

Mr. A. Kyle Wing, Jr., Chairman, Fed. Telecommunication Labs., 500 Washington Ave., Nutley 10, N.J.

29 Transactions, *Vol. ED-1, No. 3-4; ED-3, No. 2-4; ED-4, No 2-3, 4; ED-5, No. 2, 3, 4; ED-6, No. 1, 3; ED-7, No. 1, 2, 3.

Electronic Computers

Annual fee: $4.

Design and operation of electronic computers.

Dr. A. A. Cohen, Chairman, Remington-Rand Univac, St. Paul 16, Minn.

35 Transactions, EC-6, No. 2, 3; EC-7, No. 1, 2, 3, 4; EC-8, No. 1, 2, 3, 4; EC-9, No. 1, 2, 3.

Engineering Management

Annual fee: $3.

Engineering management and administration as applied to technical, industrial and educational activities in the field of electronics.

Dr. Henry M. O'Bryan, General Tel. & Elec. Lab., 730 3rd Ave., New York 17, N.Y.

18 Transactions, EM-3, No. 2, 3; EM-4, No. 1, 3, 4; EM-5, No. 1-4; EM-6, No. 1, 2, 3; EM-7, No. 1, 2.

Engineering Writing and Speech

Annual fee: $2.

The promotion, study, development, and improvement of the techniques of preparation, organization, processing, editing, and delivery of any form of information in the electronic-engineering and related fields by and to individuals and groups by means of direct or derived methods of communication.

John M. Kinn Jr., Chairman, IBM Journal, 545 Madison Ave., New, York, N.Y.

7 Transactions, Vol. EWS-1, No. 2, EWS-2, No. 1, 2, 3; EWS-3, No. 1, 2.

THE INSTITUTE OF RADIO

-IRE's 28 Professional Groups

the group chairman, and publications to date.

* Indicates publications still available

Human Factors in Electronics

Annual fee: $2.

Development and application of human factors and knowledge germane to the design of electronic equipment.

Mr. Robert R. Riesz, Chairman, Bell Tel. Labs, Murray Hill, N.J.

1 Transaction, HFE-1, No. 1.

Industrial Electronics

Annual fee: $3.

Electronics pertaining to control, treatment and measurement, specifically, in industrial processes.

Mr. J. E. Eiselein, Chairman, RCA Victor Div., Camden, N.J.

13 Transactions, *PGIE 1, 3, 5, 6, 7, 8, 9, 10, 11; IE-7, No. 1, 2.

Information Theory

Annual fee: $4.

The theoretical and experimental aspects of information transmission, processing and utilization.

Mr. Paul E. Green, Jr., Chairman, 14 Bradford Rd., Weston 93, Mass.

22 Transactions, PGIT-4, IT-1, No. 3; IT-2, No. 3; IT-3, No. 1, 2, 3, 4; IT-4, No. 1, 2, 3, 4; IT-5, No. 1, 2, 3, 4; IT-6, No. 1, 2, 3.

Instrumentation

Annual fee: $2.

Measurements and instrumentation utilizing electronic techniques.

Mr. C. W. Little, Jr., Chairman, C-Stellerator Assoc., Princeton, N.J.

17 Transactions, PGI-4, Vol. 1-6, No. 2, 3, 4; Vol. 1-7, No. 1, 2; Vol. 1-8, No. 1, 2; Vol. 1-9, No. 1.

Microwave Theory and Techniques

Annual fee: $3.

Microwave theory, microwave circuitry and techniques, microwave measurements and the generation and amplification of microwaves.

Dr. Kiyo Tomiyasu, Chairman, General Electric Gen. Eng. Lab., Schenectady, N.Y.

31 Transactions, MTT-4, No. 3; MTT-5, No. 3, 4; MTT-6, No. 1, 2, 3, 4; MTT-7, No. 2, 3, 4; MTT-8, No. 1, 2, 3, 4.

Military Electronics

Annual fee: $2.

The electronics sciences, systems, activities and services germane to the requirements of the military. Aids other Professional Groups in liaison with the military.

Dr. Edward G. Witting, Chairman, 3700 N. Albemarle St., Arlington 7, Va.

9 Transactions, MIL-1, No. 1; MIL-2, No. 1; MIL-3, No. 2, 3, 4; MIL-4, No. 2-3.

Nuclear Science

Annual fee: $3.

Application of electronic techniques and devices to the nuclear field.

Mr. Louis Costrell, Chairman, 10614 Cavalier Dr., Silver Spring, Md.

17 Transactions, NS-1, No. 1; NS-4, No. 2; NS-5, No. 1, 2, 3, NS-6, No. 1, 2, 3, 4; NS-7, No. 1, 2-3.

Product Engineering & Production

Annual fee: $2.

New advances and materials applications for the improvement of production techniques, including automation techniques.

Mr. W. D. Novak, Chairman, 325 Douglas Road, Chappaqua, N.Y.

6 Transactions, No. 2-3, 4, 5, 6.

Radio Frequency Techniques

Annual fee: $2.

Origin, effect, control and measurement of radio frequency interference.

Professor Ralph M. Showers, Chairman, Moore School of Elec. Eng., 200 S. 33rd St., Philadelphia 4, Pa.

2 Transactions, RFI-1, No. 1, RFI-2, No. 1.

Reliability and Quality Control

Annual fee: $3.

Techniques of determining and controlling the quality of electronic parts and equipment during their manufacture.

Mr. P. K. McElroy, Chairman General Radio Co., West Concord, Mass.

18 Transactions, *3, 5, 10, 11, 12, 13, 14, 15; RQC-9, No. 1, 2.

Space Electronics and Telemetry

Annual fee: $2.

The control of devices and the measurement and recording of data from a remote point by radio.

Mr. Robert V. Werner, Chairman, 5575 Kearney Road, San Diego 10, Calif.

15 Transactions, TRC-1, No. 2-3; TRC-2, No. 1; TRC-3, No. 2, 3; TRC-4, No. 1; SET-5, No. 1, 2, 3, 4; SET-6, No. 1, 2.

Ultrasonics Engineering

Annual fee: $2.

Ultrasonic measurements and communications, including underwater sound, ultrasonic delay lines, and various chemical and industrial ultrasonic devices.

Mr. David L. Arenberg, Chairman, Arenberg Ultrasonic Lab., Inc., 94 Green St., Jamaica Plains, Mass.

9 Transactions, PGUE, 5, 6, 7; UE-7, No. 1, 2.

Vehicular Communications

Annual fee: $2.

Communications problems in the field of land and mobile radio services, such as public safety, public utilities, railroads, commercial and transportation, etc.

Mr. Richard P. Gifford, Chairman, 2719 Hurdle Hill Rd., Lynchburg, Va.

15 Transactions, 5, 8, 9, 10, 11, 12, 13; Vol. VC-9, No. 1, 2.

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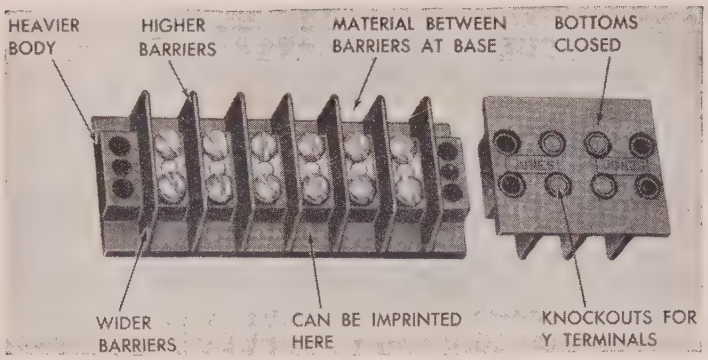

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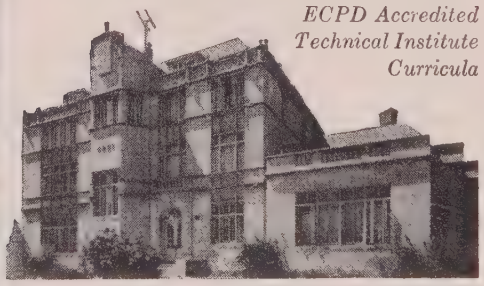

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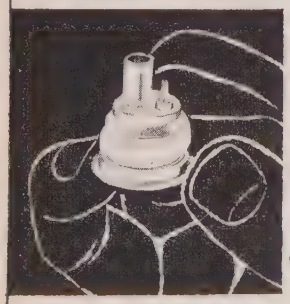
SEAL PROBLEMS? *A CASE IN POINT*



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GENERAL ELECTRIC required development of a rugged, compact high current hermetic seal CONTROLLED RECTIFIER housing constructed of materials and processes to withstand temperatures above soft solder range—design involves 5 seals to dissimilar materials.

SOLUTION:
Mechanical requirement dictated use of 3 metals: alloy #52, OFHC and Gr "A" Ni. Braze material selected is above 1435°F. so that subsequent welding or brazing can be done without detrimental effect. Tapered seals eliminate costly ground ceramics.

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IRE People

(Continued from page 58A)

conductor Division's research center at Newark, N. J.

Jerome Fishel (A'55) was named Vice President-Sales; he was formerly Vice President-Marketing of General Transistor.

❖

Dr. Elie J. Baghdady (A'52–M'57), a noted authority on modulation theory and communication systems, was recently named Director of Research of the Advanced Communications Research Division of General Electronic Laboratories, Inc., Cambridge, Mass.

He was born in Zahleh, Lebanon, on November 10, 1930. Between 1949 and 1951 he was a British Council Scholar and recipient of the first-prize scholarship at the American University of Beirut. He received the B.A. degree with high distinction in physics in 1951. In the summer of the same year he came to the U. S. on a Georgia Institute of Technology World Student Fund Scholarship, and enrolled in the Institute's graduate division of electrical engineering at Atlanta, Ga.

In 1952 he joined the electrical engineering staff of the Massachusetts Institute of Technology, Cambridge, as a graduate student and research assistant in the Research Laboratory of Electronics. He became an instructor in electrical engineering in 1955, and assistant professor in 1956. He received the S.M. and ScD. degrees in electrical engineering from M. I. T. in 1954 and 1956, respectively. In 1958 the Government of Lebanon awarded him the Lebanese Golden Medal of Distinction.

His research has been mainly concerned with communication and modulation studies, including problems of improving reception in the presence of high-level noise and interference and the analysis of linear and nonlinear system response to FM signals.

Dr. Baghdady is a member of Sigma Xi and the IRE Professional Groups on Communications Systems, Circuit Theory, Information Theory and Vehicular Communications.

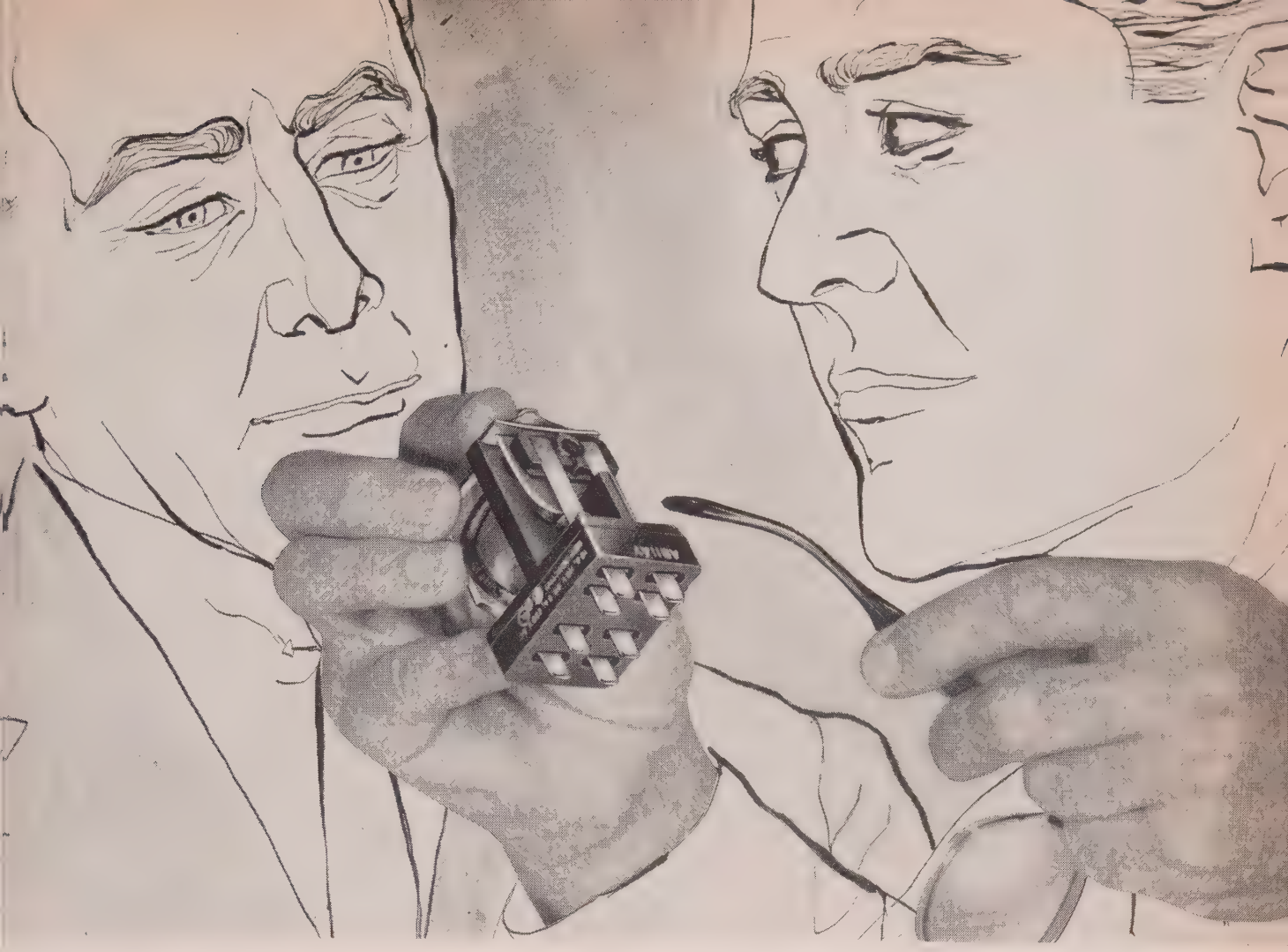
❖

George Houseworth (A'48–M'55–SM'57) was recently presented with the Commanding Officers' Certificate of Merit Award by the Commanding Officer and Director of the U. S. Naval Training Device Center, Port Washington, N. Y. He was nominated as the Center's candidate for the outstanding service award presented annually by the Federal Businessmen's Association of New York.

Mr. Houseworth was born in New York, N. Y., on February 2, 1923. He received the B.S. and M.S. degrees from Brooklyn Polytechnic Institute, Brooklyn, N. Y., and is presently employed as a General Engineer in the Special Projects Office of the Center.

❖

(Continued on page 66A)



appearances are not deceiving

THIS P&B 10-AMP RELAY IS AS RELIABLE AS IT LOOKS

Our AB relay looks rugged . . . and it is. You can specify it for 10 amp switching and confidently expect 100,000 cycles. Yet it is compact, easily mounted, and does not require special handling. Installation is simple, using your preference of screw terminals (adapters), quick connects, or dip soldering.

Designers specify the AB for air conditioners and other products where dependable, continual service is paramount.

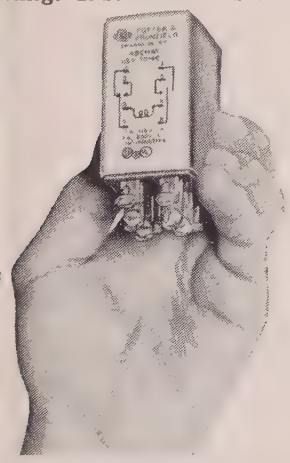
These standard AB and ABC relays are listed by Underwriters' Laboratories and Canadian Standards Association:

Type	Arrangements	Type	Arrangements
AB7AY	DPST-NO	ABC7AY	DPST-NO
AB8AY	DPST-NC	ABC8AY	DPST-NC
AB11AY	DPDT	ABC11AY	DPDT

Coil voltages: 6, 12, 24, 115 and 230 volts AC, 50/60 cycle.
Contact rating 10 amps, 115 volts AC or 5 amps, 230 volts AC noninductive.

U/L File E-29244 CSA No. 15734

Write for complete data or contact your nearest P&B sales engineer.



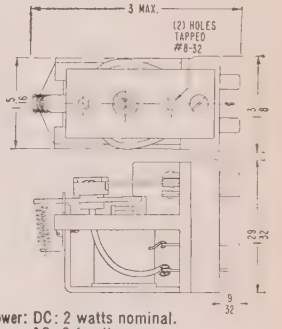
ABC Series—AB series can be supplied enclosed in sturdy metal dust cover, 1 31/64" x 2 25/32" x 2 3/32".

AB AND ABC RELAYS ENGINEERING DATA

GENERAL:
Insulation Resistance: 100 megohms minimum.
Life: 3 million cycles (mechanical).
Breakdown Voltage: 1500 volts rms between all elements and ground.
Temperature Range: DC: −55 to +45°C.
AC: −55 to +45°C.
Weight: AB—5 ozs. ABC—7 ozs.
Terminals: Fit ¼" quick-connect terminals, or may be applied to printed circuits using dip soldering. Screw adapters furnished on request.
Enclosure: ABC: Heavy duty dust cover.
Dimensions: 1 31/64" x 2 25/32" x 2 3/32".

CONTACTS:
Arrangements: DPDT
Material: ¼" dia. silver. Other materials available.
Load: 5 amps at 230 volts AC or 10 amps at 115 volts AC noninductive.
10 amps at 28 volts DC.

COIL:
Voltage: DC: 6 to 110 volts.
AC: 6 to 230 volts.



Power: DC: 2 watts nominal.
AC: 6.4 volt-amps.
Resistance: 35,000 ohms max.
Duty: Continuous: DC coils will withstand 6 watts at +25°C.

MOUNTINGS:
AB: Two 8-32 tapped holes on 1¼" centers.
ABC: One 8-32 stud ⅜" long and locating tab.

P & B STANDARD RELAYS ARE AVAILABLE AT YOUR LOCAL ELECTRONIC PARTS DISTRIBUTOR.

AMF **POTTER & BRUMFIELD**

DIVISION OF AMERICAN MACHINE & FOUNDRY COMPANY, PRINCETON, INDIANA

IN CANADA: POTTER & BRUMFIELD CANADA LTD., GUELPH, ONTARIO

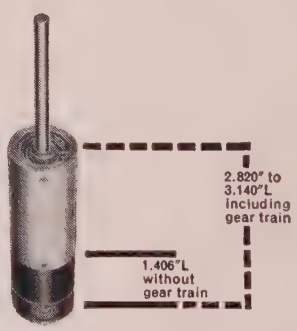
Virtually No Speed Variation From −67°F to +167°F Due To Special Dry Lubrication Method

Type 9001-03 Size 9 28VDC permanent magnet timer motor or motor tach

- Withstands 8000 G axial shock; 500 G transverse shock.
- Up to 500 oz. in. peak load can be handled.
- Designed to perform 2,000 duty cycles at 100 oz. in. load.
- ±.0001 shaft diameter tolerance.
- +.000 −.0005 shaft flat tolerance.
- 1.406″L without gear train; 2.820″ to 3.140″L including gear train.
- 6:1 to 46,656:1 gear ratios available.

GEAR RATIO	OUTPUT RPM	LOAD	AMPS AT 25°C	UNIT LENGTH	OSTER TYPE NO.
46,656 to 1	0.25	100 oz. in.	0.075 max.	3.140	E149
4536 to 1	2.68	100 oz. in.	0.100 max.	2.980	E169
756 to 1	12	100 oz. in.	0.175 max.	2.820	E170
7776 to 1	1.5	100 oz. in.	0.090 max.	2.980	E171
9168 to 1	1.25	100 oz. in.	0.085 max.	3.140	E179
4056 to 1	2.75	100 oz. in.	0.100 max.	2.980	E215
2923 to 1	4.00	100 oz. in.	0.120 max.	2.980	9001-05
393.7 to 1	33	32 oz. in.	0.200 max.	2.725	9001-02*

Max. shaft dia. .203″; .250″ on Type 9001-02. *Not a hi-shock unit. Output 27.5VDC.



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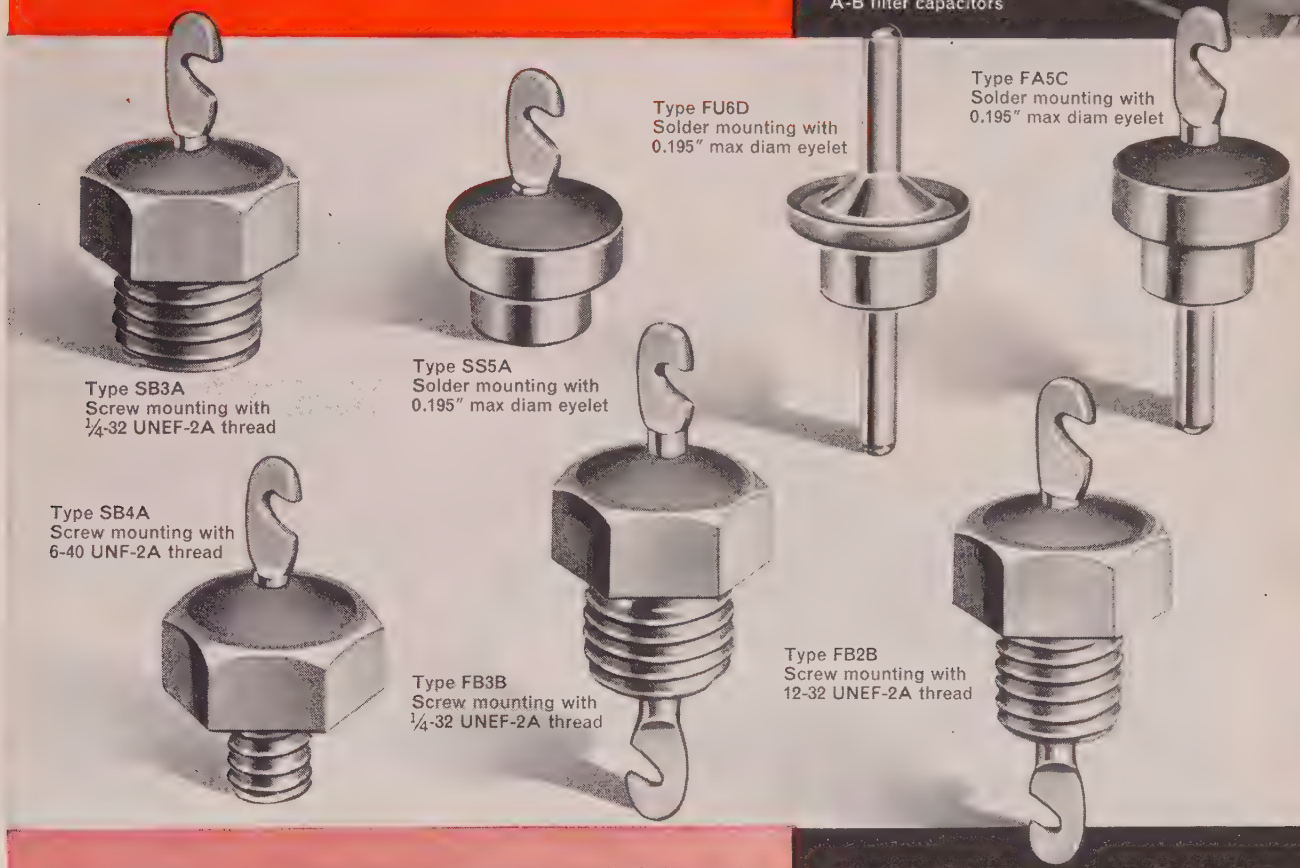
New Allen-Bradley LINE OF Discoidal Capacitors PROVIDES *SUPERIOR PERFORMANCE, PHYSICAL UNIFORMITY*

ACTUAL SIZE
of the seven standard
A-B filter capacitors



Type SB3A
Screw mounting with
¼-32 UNEF-2A thread

Type SS5A
Solder mounting with
0.195″ max diam eyelet

Type FU6D
Solder mounting with
0.195″ max diam eyelet

Type FA5C
Solder mounting with
0.195″ max diam eyelet

Type SB4A
Screw mounting with
6-40 UNF-2A thread

Type FB3B
Screw mounting with
¼-32 UNEF-2A thread

Type FB2B
Screw mounting with
12-32 UNEF-2A thread

NO PARALLEL RESONANCE EFFECTS AT 1000 MCPS OR LESS. The exclusive dielectric construction of Allen-Bradley Discoidal Feed-thru Capacitors eliminates all parallel effects normally associated with tubular feed-thru designs. This complete freedom from self-resonance permits the use of much higher capacitance values, providing far greater attentuation of undesired radiation.

Allen-Bradley Discoidal Stand-off Capacitors provide maximum capacity for by-pass purposes in a minimum of size.

These extremely compact Allen-Bradley capacitors have an unusually rugged structural design, which provides complete and permanently mechanical protection as well as ease of assembly. The gold-plated terminals assure excellent solderability.

Insist on Allen-Bradley *quality* discoidal capacitors—they have no equal on the market. Write today for Technical Bulletin 5409.

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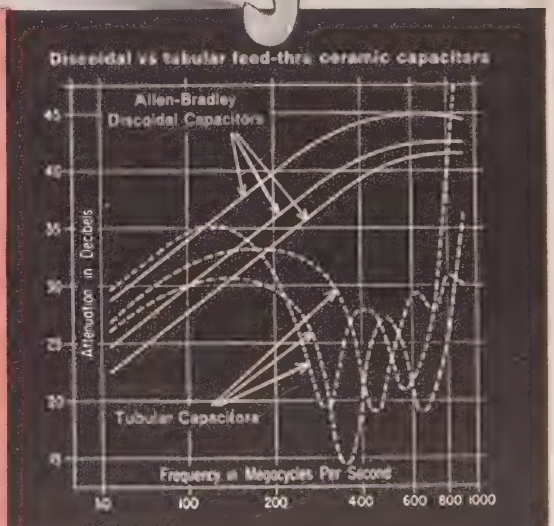
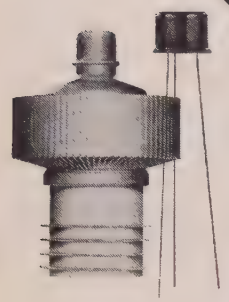
ALLEN-BRADLEY



QUALITY ELECTRONIC COMPONENTS



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with the plus quality you need for vacuum tube envelopes and internal spacers, rf windows, heat sinks, high-voltage insulators, semiconductor components and a wide array of precision electronic applications.

Wesgo's specialization in ceramics for electronics gives you unique reliability in components you can depend on.

Three dense, vacuum-tight Wesgo alumina bodies, with up to 99.5% Al_2O_3, are strong, hard and abrasion resistant. They offer exceptional chemical inertness, high thermal conductivity and superior electrical properties, even at extremely high temperatures.

All Wesgo high alumina ceramics feature very low loss factor, high dielectric strength and high resistivity. All are available in sizes and shapes to meet your individual specifications. Manufacturing is to tight dimensional tolerances; ceramics are of uniform density, free from internal and surface defects. All are quality-controlled to meet unparalleled performance standards, with extraordinary care taken in each step of the manufacturing process.

Also available from Wesgo, a complete line of low vapor pressure, high purity, precious metal brazing alloys. Write today for information

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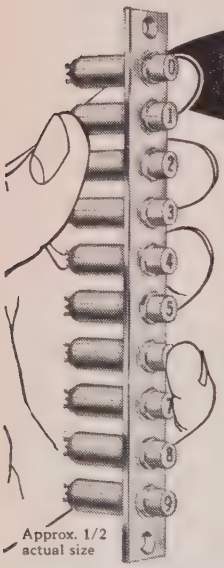

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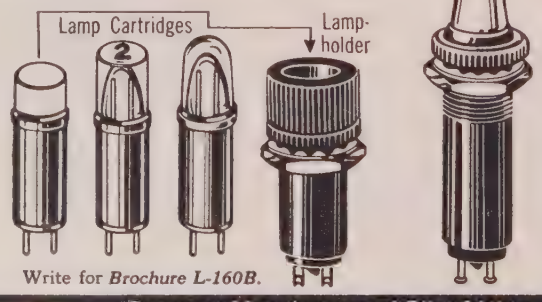
DATALITES by DIALCO are ultra-miniature Indicator Lights, made in 2 basic styles: *Lampholders* with DIALCO'S own *replaceable Lamp Cartridges;* or integrated DATALITES *with Built-in Neon Lamps** ...Mount in ⅜" clearance hole...LAMPS USED: T-1¾ Incandescent; also Neon NE-2E or NE-2H.

Approx. 1/2 actual size



DATA STRIP No. DSV-7538-10
Vertical ... complete with ten No. 39-28-1475 Lamp Cartridges. Other configurations to order.

Shown actual size, left to right: **Lamp Cartridges**—Nos. 39-6-1471, 38-1531, 38-931 ... **Lampholder** No. 7538 ... **Datalite** No. 249-7841-931 with built-in Neon Lamp and resistor.*

Write for *Brochure L-160B.*

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"The Eyes of Your Equipment"

Foremost Manufacturer of Pilot Lights

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IRE People

(Continued from page 62A)

Dr. Richard P. Lagerstrom (S'52-M'57) has joined Eitel-McCullough, Inc., San Carlos, Calif., as a senior scientist in the company's Research Division.

R. P. LAGERSTROM

As a Research Assistant and Associate with Stanford University's Electronics Research Laboratories since 1951, and with Hughes Aircraft Company from 1946 to 1951, he has worked on the development of various new techniques and methods in the field of traveling-wave tubes and microwave devices. He has been engaged in plasma dynamics studies since 1958.

Dr. Lagerstrom received the B.S. degree in electrical engineering from the California Institute of Technology, Pasadena, and the M.S. and Ph.D. degrees in the same field from Stanford University, Stanford, Calif. He is a member of the American Physical Society and has authored various technical papers.

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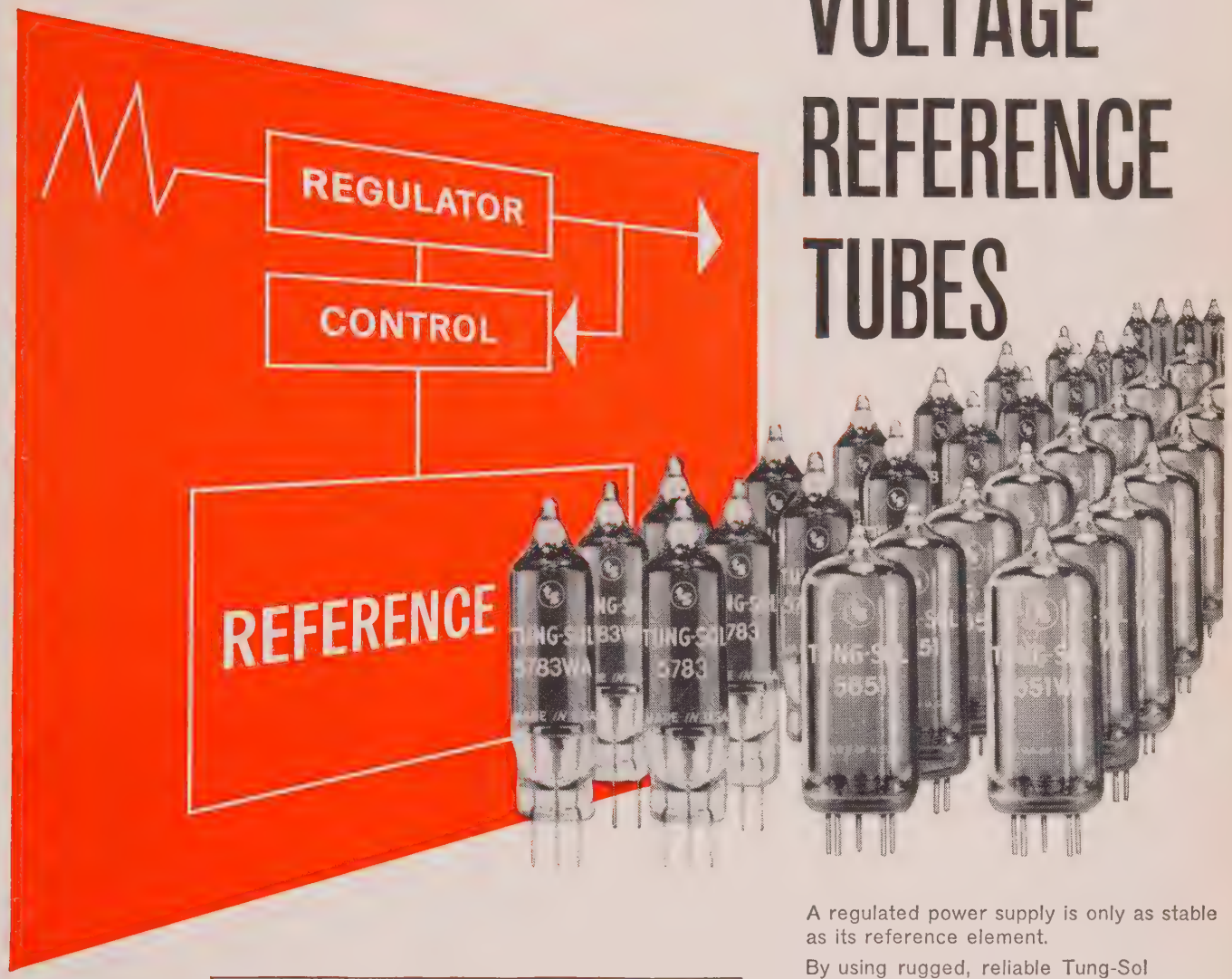
(Continued on page 68A)

Reliable stability for regulated power supplies

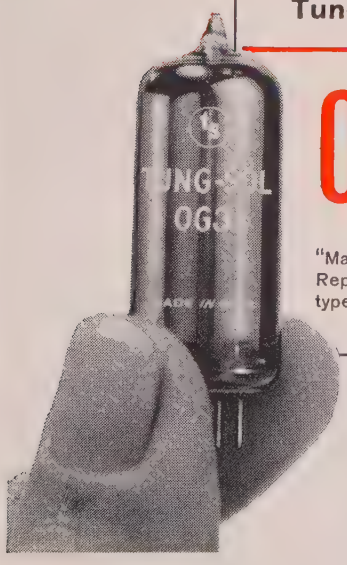
TUNG-SOL VOLTAGE REFERENCE TUBES



A regulated power supply is only as stable as its reference element.

By using rugged, reliable Tung-Sol miniature and sub-miniature reference tubes in regulated power supply designs, you're guaranteed long-life stability in spite of temperature changes, repetitive on-off cycles and current shifts. Long-term drift and noise are minimized by superior tube design, extensive factory stabilizing schedules and severe test criteria. These inert-gas-filled, cold cathode glow discharge diodes feature high shock and vibration ratings. They are particularly suited for applications where size and weight must be held to a minimum. Write for complete technical details. Tung-Sol Electric Inc., Newark 4, N. J. TWX: NK193.

Newest Member of Tung-Sol Reference Tube Family



OG3

"Made In USA"
Replaces European type 85A2

Nominal Operating Voltage—85 volts.

Minimum Operating Current—1 mAdc.

Maximum Operating Current—10 mAdc.

TUNG-SOL®

Technical Assistance is available from the following sales offices: Atlanta, Ga.; Columbus, Ohio; Culver City, Calif.; Dallas, Texas; Denver Colo.; Detroit, Mich.; Irvington, N. J.; Melrose Park, Ill.; Newark, N. J.; Philadelphia, Pa.; Seattle, Wash. Canada: Toronto, Ont.

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HIGH PERFORMANCE
Marconi microwave systems, with capacities from 60 to 960 channels, and capable of carrying high quality television, are designed to meet exacting international standards of performance with margins in hand.

EXTREME SIMPLICITY
Travelling wave tube techniques ensure extremely simple circuitry and make full use of high gain and great band width available. A unidirectional repeater consists of only three travelling wave tube amplifiers and one frequency change oscillator with their power supplies.

GREAT RELIABILITY
The use of travelling wave tubes in the repeaters has allowed considerable reduction in the number of valves and components used. Thus the likelihood of unexpected failure has been considerably reduced.

EASY MAINTENANCE
The design of the units ensures easy access to all parts of the equipment and the extensive use of printed circuitry allows speedy and accurate replacement of precision circuits by technician staff, without realignment of the equipment.

EXTREME SAFETY
All high voltages are fully interlocked.

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H 1C

IRE People

(Continued from page 66A)

Myron D. Lockwood (M'46) has been appointed vice president of Sperry Gyroscope Company, Division of Sperry Rand Corporation, Great Neck, N. Y. Mr. Lockwood, who joined Sperry as a project engineer in 1945, has been manager of the Surface-Armament Division since 1957.

M. D. Lockwood

He graduated from Virginia Polytechnic Institute in 1934 with the B.S.E.E. degree. During World War II he served with the U. S. Army and achieved the rank of lieutenant colonel. He was at one time technical advisor at the Electronics Training Center at Massachusetts Institute of Technology.

Mr. Lockwood is a member of the AIEE, Omicron Delta Kappa, and Phi Kappa Phi.

❖

The appointment of **Allen J. Lovenstein** (SM'57) as manager of market planning and analysis for the Waltham Laboratories of Sylvania Electronic Systems, a division of Sylvania Electric Products Inc., was recently announced.

A. J. Lovenstein

He previously was project engineer for the Crosley Division of Avco Corporation in charge of development of a steerable array communication system.

In his new capacity, he will prepare long-range marketing plans and sales forecasts and recommend marketing and engineering programs for the laboratories.

He served with the U. S. Army Signal Corps during World War II as a member of an American liaison team with the British and Nationalist Chinese armies in Burma.

In 1947, he joined the Signal Corps radar division engineering laboratories at Fort Monmouth, N.J., where he served until 1956 as project engineer on a series of task forces directly related to the defense effort.

Mr. Lovenstein received the bachelor's degree in electrical engineering from City College of New York, N. Y.

He is a member of the Armed Forces Communications and Electronics Association and the Electronic Industries Association.

❖

(Continued on page 74A)

Reliability in volume...
CLEVITE
TRANSISTOR
WALTHAM, MASSACHUSETTS

NEW CIRCUIT POSSIBILITIES

for low impedance, high current applications

SILICON SWITCHING DIODES

Combining high reverse voltage, high forward conductance, fast switching and high temperature operation, these diodes approach the ideal multi-purpose device sought by designers; they open new areas of opportunity for circuit design.

Type CSD-2542, for example, switches from 30 ma to −35v. in 0.5 microseconds in a modified IBM Y circuit and has a forward conductance of 100 ma minimum at 1 volt.

GENERAL PURPOSE TYPES

Optimum rectification efficiency rather than rate of switching has been built into these silicon diodes. They feature very high forward conductance and low reverse current. These diodes find their principal use in various instrumentation applications where the accuracy or reproducibility of performance of the circuit requires a diode of negligible reverse current. In this line of general purpose types Clevite has available, in addition to the JAN types listed below, commercial diodes of the 1N482 series.

MILITARY TYPES

JAN		SIGNAL CORPS	
1N457	MIL-E-1/1026	1N662	MIL-E-1/1139
1N458	MIL-E-1/1027	1N663	MIL-E-1/1140
1N459	MIL-E-1/1028	1N658	MIL-E-1/1160
		1N643	MIL-E-1/1171

Write for Bulletins B217A-1, B217A-2 and B217-4.

A DIVISION OF CLEVITE CORPORATION

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200-Ampere Collector Curves

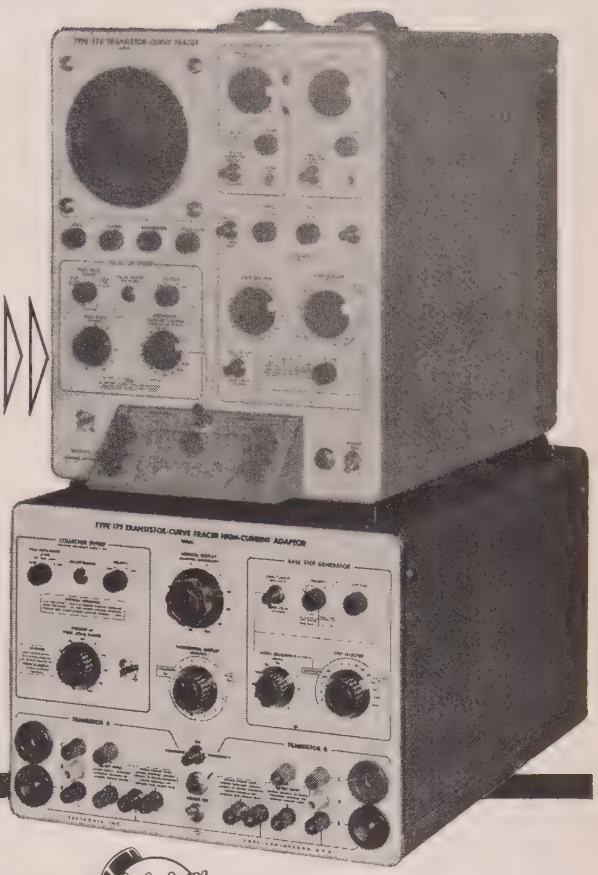
... with New High-Current Adapter for your Tektronix Transistor-Curve Tracer

Designed specifically for use with a Tektronix Type 575 Transistor-Curve Tracer, the Type 175 provides 200-ampere collector displays, three ranges of collector supply, and 12-ampere base supply. It enables the Transistor-Curve Tracer to plot and display on its crt the characteristic curves of high-powered transistors. Not intended for separate use, the Type 175 depends upon circuitry in the Type 575 for proper operation.

The Tektronix Type 575 Transistor-Curve Tracer supplies the number of steps per family, the number of steps per second, and the function of repetitive displays or single family presentations. It also provides crt positioning and deflection voltages. A connector must be added to the rear of the Type 575 which accepts a single multi-conductor interconnecting cable through which the proper signals between instruments are transferred. (This connector is factory installed above serial number 2827.)

Used with a Type 575, the Type 175 Adapter allows observation and measurement of characteristic curves of both NPN and PNP transistors, and diodes. It permits display of characteristic curves with collector current on the vertical axis and either collector-to-emitter voltage or base-to-emitter voltage on the horizontal axis. Equal steps of current, or equal steps of voltage are applied to the transistor input. For each input step, the voltage applied to the collector is swept from zero to a selected value. Generally, for transistors, a family of collector-current curves can be plotted to 200 amperes or more. (For diodes, curves can be plotted to 100 amperes or more.)

Two transistors can be compared easily by switching the test conditions from one to the other.



PERFORMANCE CHARACTERISTICS

200-AMPERE COLLECTOR DISPLAYS (*Single Family Presentation*)

100 ampere peak continuous supply current.

1-kw continuous collector power available.

3 RANGE COLLECTOR SUPPLY (Plus or Minus)

0 to 20 volts—0.03 Ω plus current sampling resistance.

0 to 100 volts—0.5 Ω plus current sampling resistance. (Voltage drop across current sampling resistance—0.1 v/div.)

0 to 100 volts—300 Ω series load resistor.

12-AMPERE BASE SUPPLY

Positive or negative base stepping (4 to 12 steps per family, either repetitive or single family presentation, from Type 575).

10 current step positions ranging from 1 to 1000 milliamperes per step.

5 voltage step positions ranging from 0.02 to 0.5 volts per step—with 11 different driving resistances.

A switch permits insertion of any one of ten resistances (from 0.5 to 1000 ohms) in series with the voltage step generator.

Step zero control allows adjustment of the starting point on the output current or voltage step.

A switch permits either a zero current or zero voltage reference check.

CALIBRATED DISPLAYS

Vertical Axis—Collector Current in 12 steps ranging from 0.005 to 20 amperes per division.

Horizontal Axis—Collector Voltage (V_{ce}) in 7 steps ranging from 0.1 to 10 volts per division.

Base Voltage (V_{be}) in 5 steps ranging from 0.1 to 2 volts per division.

4-TERMINAL VOLTAGE SENSING FACILITY

The voltage at the collector and the emitter of either transistor under test may be sampled at the transistor terminals.

MECHANICAL FEATURES

VENTILATION—Filtered, forced-air circulation maintains safe operating temperature.

CONSTRUCTION—Compact unit with aluminum-alloy chassis and cabinet. Side panels and bottom panel are easily removable. Components are readily accessible.

FINISH—Photo-etched anodized front panel, blue vinyl-finish cabinet.

DIMENSIONS—9¾" high by 13½" wide by 24" long.

WEIGHT—95 pounds.

POWER REQUIREMENTS—Operates from 105 to 125 volts, 50 to 60 cycles.

TYPE 175 Transistor-Curve Tracer High-Current Adapter **$1425**

. . . for operation from 210 to 250 volts, 50 to 60 cycles, please order **Type 175** MOD 161C **$1425**

f.o.b. factory

Tektronix, Inc.

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TEKTRONIX ENGINEERING REPRESENTATIVES: Hawthorne Electronics, Portland, Oregon • Seattle, Washington. Tektronix is represented in twenty overseas countries by qualified engineering organizations

In Europe please write Tektronix Inc., Victoria Ave., St. Sampsons, Guernsey C.I., for the address of the Tektronix Representative in your country.

How to draw on The Zeus Bank

Today you can draw on the same company that developed the complex million-dollar capacitors of the Zeus Bank* for any and every capacitor problem you face. Leader and pioneer in electronic components for half a century, Cornell-Dubilier has on file tens of thousands of capacitor specifications; is ready to develop a complete new one for you on request. Now an integral part of fast-growing Federal Pacific Electric Company.

*For thermonuclear fusion experiments at Los Alamos, N. M.

CAPACITORS: TANTALUM, MICA, CERAMIC, ELECTROLYTIC, PAPER, OIL, MYLAR, METALLIZED, ENERGY STORAGE AND POWER FACTOR CORRECTION

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CDE offers more than ever before—in plant and personnel resources, in depth of "know-how" throughout every phase of electric and electronic design, in a vigorous spirit of progress that welcomes every inquiry. Deposit your capacitor problems at CDE's door . . . you'll get assistance you can bank on, with richer dividends in creative product thinking that assures success. Cornell-Dubilier Electronics Division, South Plainfield, New Jersey.

CORNELL-DUBILIER ELECTRONICS DIV. CDE FEDERAL PACIFIC ELECTRIC COMPANY

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for use in conjunction with
Semi-Conductor Products
Platinum Alloy Ribbon
Tungsten-Platinum and Iridium-Platinum for "S" contacts on glass-sealed Silicon diodes. One surface is roughened to improve contact characteristics . . . Exceptionally good spring characteristics of 8% Tungsten-Platinum permit use of thinner ribbons, with savings in cost . . . Tungsten-Platinum withstands high sealing-in temperatures with little, if any, loss in spring characteristics.
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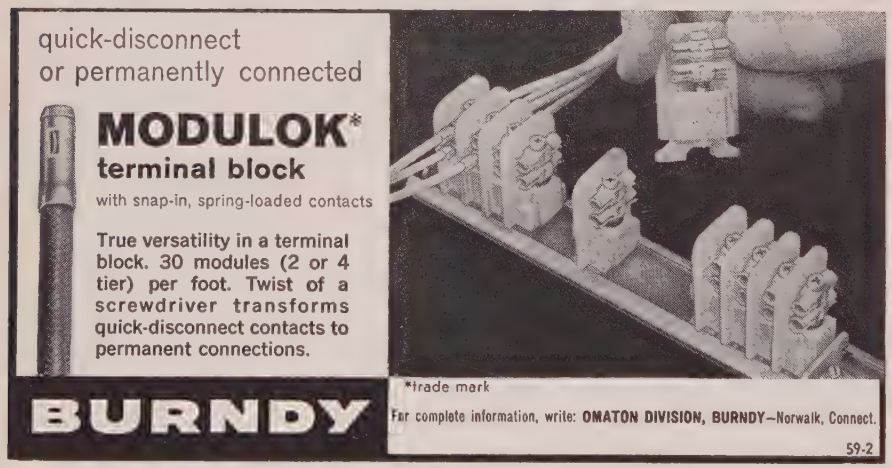
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or permanently connected
MODULOK*
terminal block
with snap-in, spring-loaded contacts
True versatility in a terminal block. 30 modules (2 or 4 tier) per foot. Twist of a screwdriver transforms quick-disconnect contacts to permanent connections.
BURNDY
*trade mark
For complete information, write: OMATON DIVISION, BURNDY—Norwalk, Connect.
59-2

IRE People

(Continued from page 68A)

William R. Luebke (S'49–A'51–M'56–SM'57) has joined the Research Division of Eitel-McCullough, Inc., San Carlos, Calif., manufacturer of Eimac electron-power tubes, as a senior scientist. As a research associate at the Electronics Research Laboratories of Stanford University from 1950 to 1960 he worked on low and medium CW, pulsed traveling-wave tubes, backward-wave oscillators, electron-beam focusing, and depressed collector studies. He was a consultant to the General Electric Microwave Laboratory at Palo Alto, Calif. from 1957 to 1960 in the field of medium-power traveling-wave tubes. He also has served as a consultant to other electronics industries.

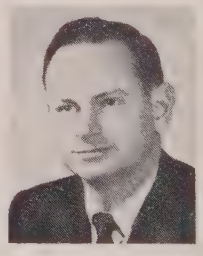
W. R. Luebke

Mr. Luebke holds the B.S. and M.S. degrees in electrical engineering from the University of Colorado.

❖

George L. Mallory (A'54–M'55) has been appointed to the newly created post of vice president—operations for Gates Radio Company, Quincy, Ill.

He was formerly manager of manufacturing for the Electronic Section of The Martin Company's Denver Missile Division, and before that, was assistant general manager of the Marine Products Division of The Bendix Corporation in Los Angeles, Calif. He is an electrical engineering graduate of Auburn University, and holds both a mechanical engineering degree and a master's degree in business administration from the University of California.



G. L. Mallory

Mr. Mallory will have over-all responsibility for manufacturing, financial, purchasing and personnel activities at Gates Radio, which operates two plants in Quincy.

❖

E. Kenneth Peterson (S'47–SM'56), formerly Chief Engineer of the Microwave Products Section of Motorola Incorporated in Chicago, has joined Lenkurt Electric Co., Inc., San Carlos, Calif.

A Motorola employee for nine and one half years, he will augment the staff of Vice President Gerd Wallenstein, Product Planning, at Lenkurt.

Mr. Peterson received the Bachelor of Science degree in Electrical Engineering

(Continued on page 78A)

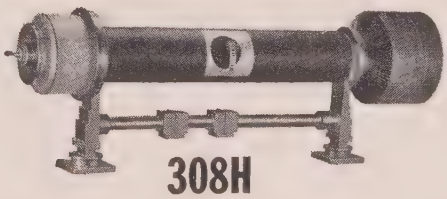
now 100 times greater average power

Now you can obtain traveling-wave tubes capable of 10 to 100 times the average power of conventional helix tubes. These X-band tubes are representative of a wide variety of the first commercially available all metal filter-type structures yielding both high gain and wide bandwidth. Their attractively small size and weight are made possible through application of the latest periodic focusing techniques.

Typical of these recent advances is the pictured 308H. For the first time a power traveling-wave tube is offered with a high-μ grid-controlled gun. This advantage, coupled with 53 db of saturation gain, provides exciting possibilities for the systems designer.

Consult with Hughes Microwave Tube Division if you have exacting design requirements of pulse rise time, phase shift sensitivity, bandwidth or power output. These qualities are yours in a light, compact, yet rugged package of all metal-ceramic construction. These advanced products can make your program a success. Orders are being accepted now for delivery in three or four months.

307H LEFT: 100 kw peak power output (500 watt average), 8.5—9.5 kmc frequency range, 54 db saturation gain, 1% maximum duty cycle, beam voltage = 38 kv, 21 lbs. total weight of tube and magnet.



308H

Control grid μ = 55. 15 kw peak power output (150 watt average), 8.6—9.9 kmc frequency range, 53 db saturation gain, 1% maximum duty cycle, beam voltage = 24 kv, 14 lbs. total weight of tube and magnet.

319H

20 kw peak power output (200 watt average), 8.4—9.6 kmc frequency range, 54 db saturation gain, 1% maximum duty cycle, beam voltage = 24 kv, 17 lbs. total weight of tube and magnet.

For full details on these and other equally outstanding tubes write or wire Hughes Microwave Tube Division, 11105 Anza Avenue, Los Angeles 45, California.

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MICROWAVE TUBE DIVISION
HUGHES AIRCRAFT COMPANY

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4 TRANSITRON TYPES AUGMENT

Silicon Controlled Rectifiers / Switches

NEW! CONTROLLED SWITCHES

TSW31S - TSW201S PNPN bistable switching devices in TO-18 packages, with maximum holding current of 1 ma.

- High gate sensitivity 20 μa to fire
- Covers current range from 1 ma to 200 ma @ 75°C ambient
- Voltage ratings up to 200 volts available
- Temperature range: −65°C to +150°C

NEW! TO-5 PACKAGE CONTROLLED RECTIFIERS

Two series of diffused silicon PNPN bistable switching devices with very low triggering requirements and micro-second switching.

TCR251-TCR4001 series featuring:

- Low leakage: 100 μA @ 125°C case
- High gate sensitivity: 200 μA @ 25°C
- Low Holding Current: 5 mA maximum @ 25°C
- Current rating: 1 amp at 80°C case or 600 ma at 25°C ambient
- Voltage ratings: Up to 400 volts

Plus **2N1595-2N1599** series with same current and voltage ratings

Series	Package
TCR 550	1⅛ hex base
TCR 520	11/16 hex base
TCR 510	11/16 hex base
TCR 505	7/16 hex base

Average Forward Current Ratings

25ma 50ma 5A 10A 20A 50A

2N1600	7/16 hex base
2N1595	TO-5
TCR 251	TO-5
TSW 31S	TO-18

(1) Maximum Holding Current Values

Average Forward Current Ratings

1ma 5ma 25ma 50ma 200ma 1A 3A

The complete Transitron line of Controlled Rectifiers and Controlled Switches includes the following higher current types:

NEW! 2N1600-2N1604 and TCR505-TCR4005 series diffused Silicon Controlled Rectifiers

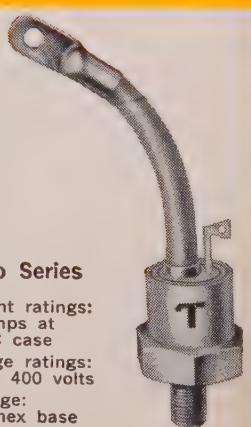
- Current ratings: 3 amps at 80°C case; 1 amp @ 125°C case
- Voltage ratings: Up to 400 volts
- Package: 7/16" hex base

10 Amp Series

- Current ratings: 10 amps @ 25°C case; 5 amps at 100°C case
- Voltage ratings: Up to 400 volts
- Package: 11/16" hex base

20 Amp Series

- Current ratings: 20 amps @ 25°C case; 10 amps at 100°C case
- Voltage ratings: Up to 400 volts
- Package: 11/16" hex base



NEW! 50 Amp Series

- Current ratings: 50 amps at 100°C case
- Voltage ratings: Up to 400 volts
- Package: 1⅛" hex base

RECTIFIERS & SWITCHES

NDUSTRY'S BROADEST LINE!

Binistors / Transwitches

THE BINISTOR

(bý-nis-tor)

Transitron's new silicon NPN Tetrode offers simpler, more reliable, more economical switching and storage circuitry. The key parameters of this bistable, negative resistance device are determined by external circuitry, providing remarkable stability and uniformity over wide temperature ranges. The signal and output swings are compatible with present transistor and diode circuits. Two series are available: The wide temperature range or military types and the commercial and industrial computer types. The stability and uniformity of each unit in the military series is absolutely guaranteed by the method of specification at critical temperatures (−65°C and +150°C).

ABSOLUTE MAXIMUM RATINGS

	3N56	3N57
Collector to Emitter Voltage (V_{CE})	15 Volts	15 Volts
Collector Current @ 25°C (I_C)	30 mA	30 mA
Storage & Operating Ambient Temp. Range	−65°C to +150°C	−55°C to +100°C

3N56 MILITARY TYPE

SPECIFICATIONS & TYPICAL CHARACTERISTICS (At Noted Ambient Temp.)

TURN-ON	AMBIENT TEMP	MIN.	TYPICAL	MAX.	TEST CONDITIONS
D.C. Collector Saturation Voltage (V_{CE})	−65°C	—	0.46	1.0 V	I_C = 10mA, I_b = + .5mA*
	+25°C	—	0.7	1.0 V	V_J = 4V, R_J = 3K
	+150°C	—	1.2	1.5 V	supply
Critical Injector Current ($I_{i\ crit}$)	−65°C	0	.38	.5 mA	I_C = 10mA, I_b = − 50μa
	+25°C	0	.28	.5 mA	
	+150°C	0	.21	.5 mA	
	−65°C	0	1.10	1.5 mA	I_C = 0.25mA, I_b = − 50μa
	+25°C	0	0.86	1.5 mA	
	+150°C	0	0.40	1.5 mA	
TURN-OFF Base Cutoff Current (I_{bo})	−65°C	—	—	—	V_{CE} = 15 volts, V_{JE} = + 13 volts
	+25°C	—	.020	0.2 μA	V_{BE} = − .6 volts
	+150°C	—	2.0	10.0 μA	

3N57 COMPUTER TYPE

SPECIFICATIONS & TYPICAL CHARACTERISTICS @ 25°C

TURN-ON	MIN.	TYPICAL	MAX.	TEST CONDITIONS
D.C. Collector Saturation Voltage (V_{CE})	—	0.7	1.0 V	I_C = 10mA, I_b = + .5mA* V_J = 4V, R_J = 3K supply
Critical Injector Current ($I_{J\ crit}$)	0	.28	0.5 mA	I_C = 10mA, I_b = − 50μa
	0	.86	1.5 mA	I_C = 0.25mA, I_b = − 50μa
TURN-OFF Base Cutoff Current (I_{bo})	—	.020	.2 μA	V_{CE} = 15 volts, V_{JE} = + 13 volts V_{BE} = − .6 volts

*Unit must switch on under the above conditions; however, actual V_{CE} measurement is made with I_b = − 50μa

THE TRANSWITCH

A PNPN bistable silicon computer element that can be turned on and off with gate current. The device is available in the TO-18 package, and is designed for miniaturized memory circuits, ring counters, shift registers, controlled rectifier drivers, and flip flop equivalents. A 100 ma series (TSW-31A-TSW-201A) has been added to the Transwitch series. Both series (50mA and 100mA) are available in voltage ratings up to 200 volts. For commercial and industrial applications, the SW-30 type is now available. This unit, especially designed for lower temperature applications, features maximum collector current rating of 30mA and maximum voltage rating 30 volts.

ABSOLUTE MAXIMUM RATINGS

	SW-30	TSW-31 thru TSW-201	TSW-31A thru TSW-201A
Forward current I_f	30 mA	50 mA	100 mA
Operating temp. range	−55°C to +85°C	−55°C to +125°C	−55°C to +125°C

SPECIFICATIONS (AT 25°C)

	SW-30	TSW-31 thru TSW-201	TSW-31A thru TSW-201A
Max. Saturation Voltage (V_S)	1.5 V @ 30 mA	1.5 V @ 50 mA	2 V @ 100 mA
Max. Forward "OFF" Current (I_{CGO})	10 μA	10 μA	10 μA
Max. Reverse Current (I_R)	10 μA	10 μA	10 μA
Max. Forward "OFF" Current (I_{CGO})	50 μA @ 85°C	50 μA @ 125°C	50 μA @ 125°C
Max. Reverse Current (I_R)	50 μA @ 85°C	50 μA @ 125°C	50 μA @ 125°C
Max. Gate Voltage to Switch "ON" ($V_{G\ ON}$)	1.0 V	1.0 V	1.0 V
Max. Gate Current to Switch "ON" ($I_{G\ ON}$)	1.5 mA	1.0 mA	1.0 mA
Max. Gate Voltage to Switch "OFF" ($V_{G\ OFF}$)	−5.0 V	−4.0 V	−6 V
Max. Gate Current to Switch "OFF" ($I_{G\ OFF}$)	−8.0 mA	−10 mA	−20 mA
Max. Holding Current (I_H)	10.0 mA	5.0 mA	7.0 mA

In writing for further information on all these devices, refer to the following bulletin numbers:

Controlled Rectifiers & Switches		Binistor & Transwitch	
TSW-31S series	Bulletin # TE-1356E	TSW-31A	Bulletin # TE-1357B-1
TCR-251 series	Bulletin # TE-1356D	TSW-31	Bulletin # TE-1357B
2N1595 series	Bulletin # TE-1356C	SW-30	Bulletin # TE-1357E
2N1600 series	Bulletin # TE-1356B-1	3N56	Bulletin # TE-1360A
TCR-505 series	Bulletin # TE-1356B	3N57	Bulletin # TE-1360B
10 amp series	Bulletin # TE-1356A-1		
20 amp series	Bulletin # TE-1356A		
50 amp series	Bulletin # TE-1356AA		

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Sprague-developed mass production and quality-control techniques assure lowest possible cost consistent with utmost quality and reliability. Here too, complete fabrication facilities permit prompt production in a full, wide range of sizes and shapes.

Look to Sprague for today's most advanced ceramic elements — where continuing intensive research promises new material with many properties extended beyond present limits.

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THE MARK OF RELIABILITY

IRE People

(Continued from page 74A)

from the University of Minnesota, Minneapolis, in 1948. He is a member of the Eta Kappa Nu honorary society.

❖

Industro Transistor Corporation, Long Island City, N. Y., has named **Robert W. Pike** (M'57) as Chief Engineer in Charge of Research and Development. He will head the company's new Semiconductor Research and Development Center in Natick, Mass.

He joined Industro July, 1960, coming from Clevite Transistor Products where he was Senior Development Engineer. While at Clevite, he developed processes for germanium *p-n-p* alloy junction transistors, silicon rectifiers and silicon mesa transistors.

He has also worked for International Business Machines as an Associate Physicist, where he worked on the development of automation for the germanium *n-p-n* alloy junction line. He was Project Engineer on *p-n-p* germanium alloy diffused core driver transistors, and pioneered diffused silicon devices for IBM.

Mr. Pike, 27, graduated from the University of Maine, Orono, in 1955. He holds the B.S. degree in Engineering Physics, and has done graduate work at

(Continued on page 82A)

MODEL **4005** *with* AMBITROL V-1

CONSTANT VOLTAGE
CONSTANT CURRENT

from the

SAME TERMINALS!

$143.50

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MODEL 4005 is a 40 volt, 500 ma, regulated DC power supply incorporating AMBITROL,* a transistorized regulator permitting continuous control of voltage or current to .05% with *adjustable automatic electronic crossover* to either voltage or current regulation.

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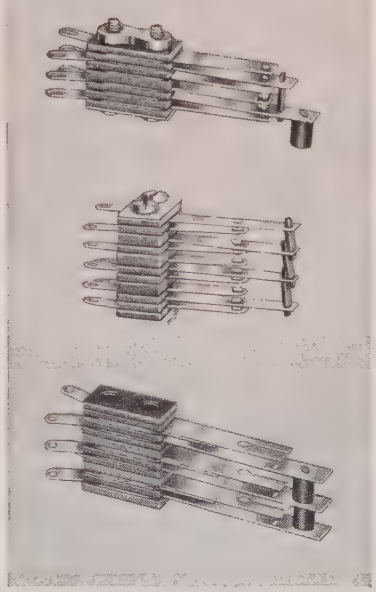
SWITCHCRAFT® INC. STACK SWITCHES



the answer to that tough Switch Problem!

.VERSATILE .DEPENDABLE .ECONOMICAL

A "pile-up" of various contact springs, insulators, contacts, etc., that can be custom-designed into your equipment. Can also be manufactured to your particular specifications.

Now being used in Tape Recorders, Remote TV Control Units, Juke Boxes, Electronic Organs, and innumerable other products requiring dependable switching.

Write for Catalog S-304. Contains a list of parts that can be made from standard tools to enable you to design a particular Switch to suit your needs.

SWITCHCRAFT® INC.

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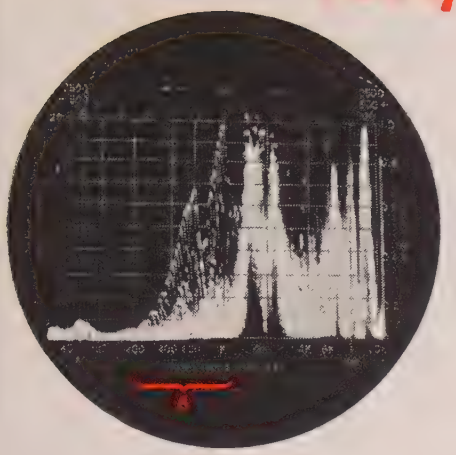
Canadian Rep.: Atlas Radio Corp., Ltd., 50 Wingold Ave., Toronto, Ontario

fast, direct-reading, easy-to-use



"Coast-down" analysis of Gyro Motor by Model LP-1a Spectrum Analyzer. Area A shows decreasing fundamental frequency, resonant rise and decay, and vibration components over 60 successive scans in one minute.

PANORAMIC LP-1a
SONIC SPECTRUM ANALYZER
20 cps-22.5 kc.

HIGHLIGHT FEATURES:

- 1-sec. "quick-look" at entire spectrum (40 cps—20 kc)
- Magnified analysis on reduced sweep widths
- Direct reading frequency-selective voltmeter
- Exceptionally stable circuitry: better than 5 cps/hr.
- Economical
- Simple Operation

The Model LP-1a "quick-look" helps locate and evaluate discrete or random signals *faster* and *easier* by scanning the entire spectrum logarithmically from 40 cps to 20 kc. Once every second it automatically separates, measures and plots the frequency and voltage of waveform components on the calibrated X and Y axes, respectively, of a long persistence 5" CRT.

For very detailed analysis, linear segments 40 to 5000 cps wide, centerable between 0 and 20 kc, may be magnified on the screen. Amplitude ratios of up to 40 db can be simultaneously measured.

High sampling rate and panoramic displays assure

1. **Minimum risk of missing signals or spectrum holes.**
2. **Fast measurements by eliminating slow point by point plots.**
3. **Simultaneous measurement of signals with widely divergent amplitudes and/or frequencies.**
4. **Continuous analysis of rapid changes in spectral content or design parameters.**

Proved in hundreds of research, design and production installations, the LP-1a is a valuable tool for Noise and Vibration analysis
• Harmonic and IM measurements • General waveform studies
• Power Spectral Density analysis • Response Curve Tracing.

SUMMARY OF SPECIFICATIONS:

Frequency Range: 20 cps—22.5 Kc.
(1) Preset linear frequency scans: any segment width of 200, 1000, 5000 cps centerable from 0-20 Kc: Variable from 40 cps to 5000 cps wtih Auxiliary Function Unit C.
(2) Preset Log Scan—40 cps to 20 Kc

Frequency Scales: Linear and Log

Center Frequency Control: Calibrated 0-20 Kc (used on lin scan)

Dynamic Range: 60 db

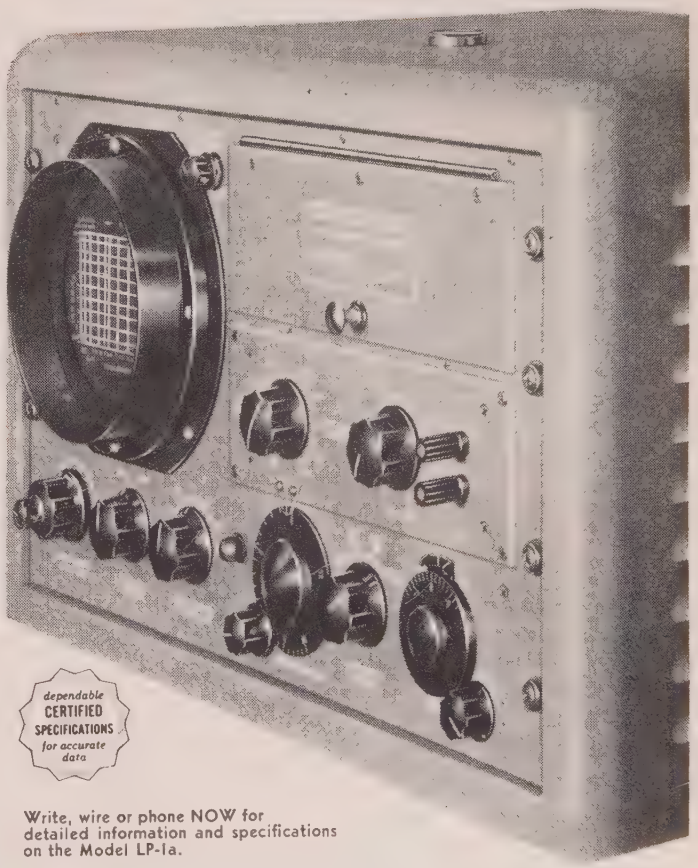
Amplitude Scales: Linear and 2 decade log (Expandable to 60 db)

Sensitivity: 500 μv to 500 v for full scale linear deflection

Voltage Accuracy: Lin Sweep (40 cps—22.5 Kc) : ±5% or ±0.5 db. Log Sweep (40 cps—20 KC) : ±10% on lin ampl. scale, ±1.5 db on log ampl. scale.

Scan Rate: 1/sec., internally generated; adjustable with accessory equipments

Resolution: For log scan, automatically optimized. For lin scan, preset 30, 75 and 170 cps at 200, 1000 and 5000 cps sweep-width, respectively. Variable from 10 cps to 1 kc with Auxiliary Function Unit C.



Write, wire or phone NOW for detailed information and specifications on the Model LP-1a.

Ask for Vols. 1, 2 and 5 of the "Panoramic Analyzer" . . . featuring applications and techniques of Spectrum Analysis, and get on our regular mailing list for this helpful publication. NEW CATALOG DIGEST AVAILABLE

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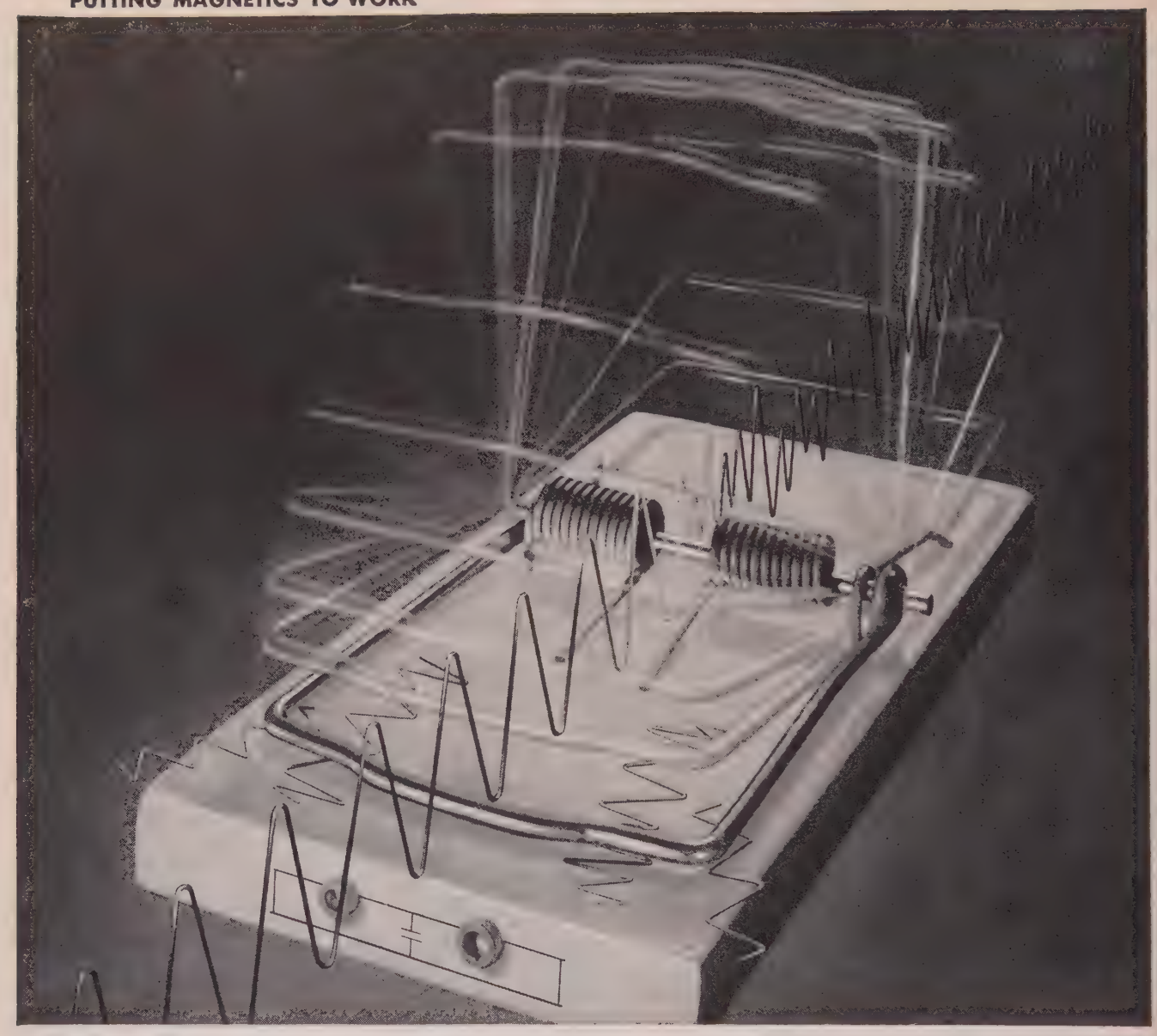
RADIO PRODUCTS, INC.



How to build a better (audio signal) trap!

Magnetics Inc. permalloy powder cores give filter designers new attenuation and stability standards—and miniaturization to boot!

The art of trapping unwanted frequencies has been advanced during the past year with a succession of improvements in molybdenum permalloy powder cores by Magnetics Inc. Most audio filter designers now work with smaller cores, more stable cores and cores whose attenuation characteristics are ultra-sharp. Do you?

Do you, for example, specify our 160-mu cores when space is a problem? With this higher inductance, you need at least 10 percent fewer turns for a given inductance than with the 125-mu core. What's more, you can use heavier wire, and thus cut down d-c resistance.

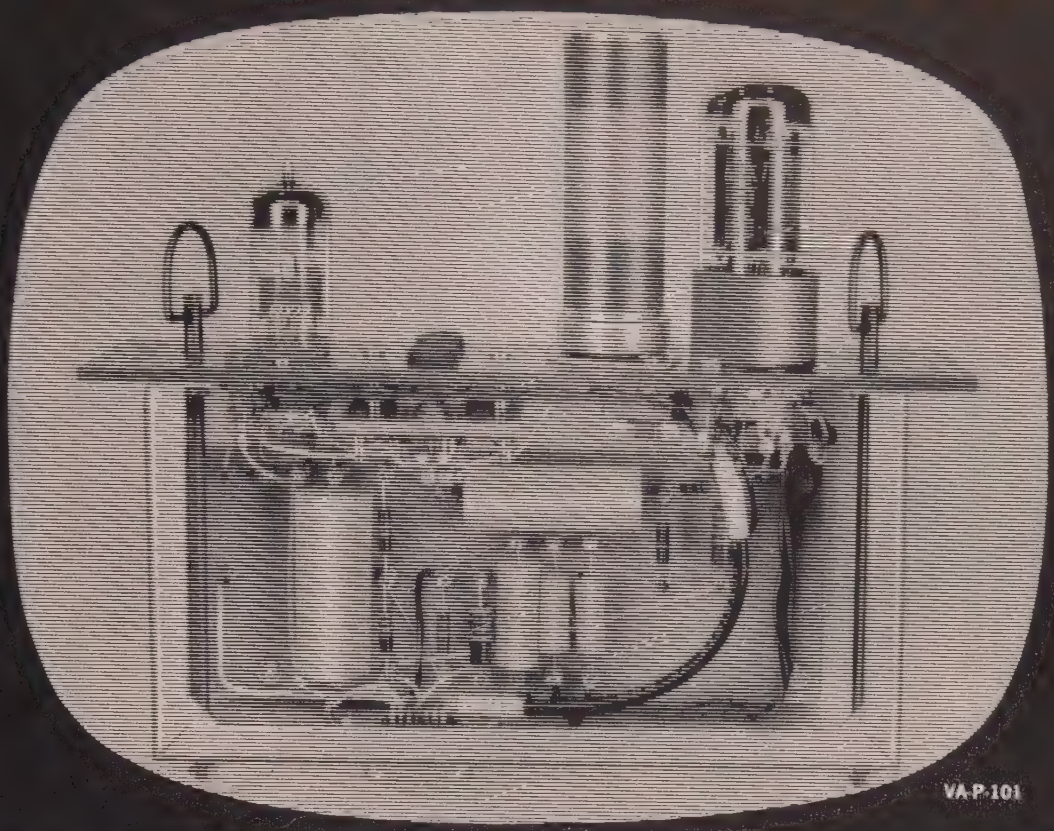
What about temperature stability? Our linear cores are used with polystyrene capacitors, cutting costs in half compared to temperature stabilized moly-permalloy cores with silvered mica capacitors. Yet frequency stability over a wide swing in ambient temperatures is increased!

And what do you specify when you must rigidly define channel cut-offs, with sharp, permanent attenuation at channel crossovers? Our moly-permalloy cores have virtually no resistive component, so there is almost no core loss. The resultant high Q means sharp attenuation of blocked frequencies in high and low band pass ranges.

Why not write for complete information? Like all of our components, molybdenum permalloy powder cores are *performance-guaranteed* to standards unsurpassed in the industry. *Magnetics Inc., Dept. P-82, Butler, Pa.*

MAGNETICS inc.®

New Daven video distribution equipment



VA-P-101

Greatest versatility for color or black-and-white!

- More outputs in less rack space
- Greatest selection to meet specific needs
- More flexibility for every possible application

Greatest versatility, combined with very low differential phase and gain, plus extremely wide band width, offers the optimum solution to all of your video distribution problems.

Plug-in Video Amplifier Specifications

Type	Description	Nominal Gain	Nominal Input Level	No. of Outputs	Bandwidth	Diff. Gain At 1V. Out	Diff. Phase At 1V. Out	Approx. B+ Drain
VA-P-101	1 in/1 out Video Dist. Amp.	Unity	1 Volt	1	Flat ±2% to at least 8.0mc	0.7% max	0.35° max	50 ma
VA-P-102	Sync Adder for VA-P-101 or VA-P-103 Amps.	—	4 Volts	1 to 7	—	—	—	30 ma
VA-P-103	1 in/1 out Video Dist. Amp.	+3db	0.7 Volt	1	Flat ±2% to at least 8.0mc	0.7% max	0.35° max	60 ma
VA-P-201-0	1 in/3 out Video Dist. Amp.	Unity	1 Volt	3	Flat ±2% to at least 8.0mc	0.7% max	0.35° max	125 ma
VA-P-201-3	1 in/3 out Video Dist. Amp.	+3db	0.7 Volt	3	Flat ±2% to at least 8.0mc	0.7% max	0.35° max	125 ma
VA-P-201-6	1 in/3 out Video Dist. Amp.	+6db	0.5 Volt	3	Flat ±2% to at least 8.0mc	0.7% max	0.35° max	125 ma
VA-P-202	Sync Adder for VA-P-201 Amps.	—	4 Volts	1 to 3	—	—	—	30 ma
VA-S-101	Mounting Shelf, to accommodate VA-P-101, VA-P-102 and VA-P-103 amplifiers. Requires 8¾ inches of rack height.							
VA-S-201	Mounting Shelf, to accommodate VA-P-201 and VA-P-202 amplifiers. Requires 8¾ inches of rack height.							

ost units are also available as "bathtub" rack mounted chassis.

Write today for new 4-page technical bulletin on Daven video equipment for color and black-and-white.

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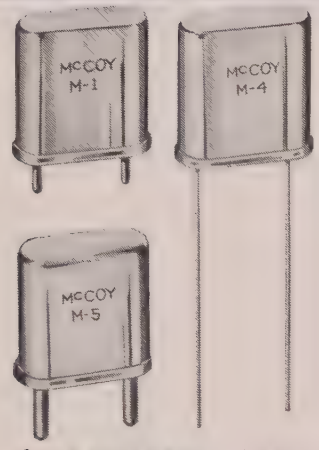
CRYSTALS and CRYSTAL FILTERS



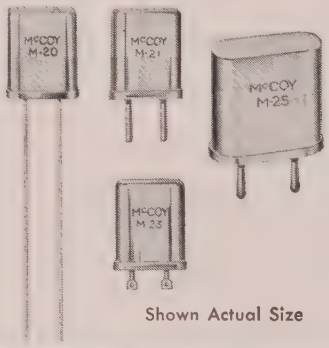
STANDARD SIZE CRYSTAL UNITS

The crystals that made the name of McCoy a synonym for quality. Metal encased, the M-1, M-4, and M-5 are available in frequencies from 500.0 kc to 200.000 mc.

Shown Actual Size

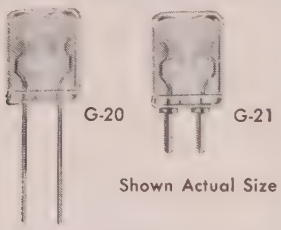
SUB-MINIATURE CRYSTAL UNITS

Fill the need for miniature crystals from 1.0 mc to 200.0 mc. Meet specs MIL-C-3098B and ARINC No. 401



Shown Actual Size

ALL-GLASS CRYSTAL UNITS



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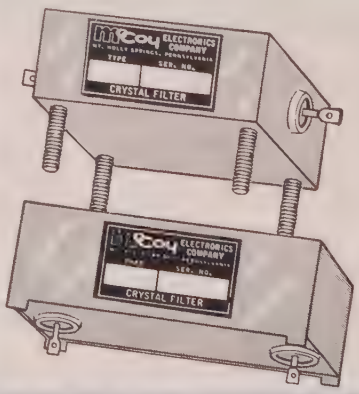
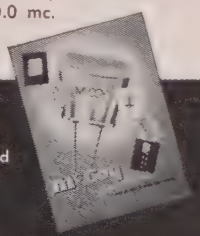
HC-18/U type. Meet new CR-73/U and CR-74/U specs. Available 5000 kc to 200.0 mc.

CRYSTAL FILTERS

Band pass types from 1.0 mc to 30.0 mc center frequency with 6 db band widths of 0.01% to 4.0% of center frequency. Single side band types from 1.0 mc to 20.0 mc frequency with 3 db bandwidths from 1.0 kc to 10.0 kc.





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IRE People

(Continued from page 78A)

Syracuse University, Syracuse, N. Y. He has also studied Business Law, Transistor Physics and Russian under the IBM Educational Program.

A native of Maine, he is a member of the Professional Group on Electron Devices, and is also a member of the Audio Society.

❖

Dr. Zvi Prihar (SM'51), former Professor of Electrical Engineering at the University of Wichita, Kan., has been appointed an Assistant Director of Telecommunications at Page Communications Engineers, Inc.

Z. PRIHAR

Prior to his academic activities, he was, during 1953–1957, Consulting Engineer and later Administrator special Studies-Communications with RCA, in their RCA International Division. His duties there included planning of telecommunications projects, special studies, foreign assignments, field surveys, and engineering consulting services. He served as Director General and Chief Engineer (Postmaster General) for the State of Israel from 1948 to 1951.

Dr. Prihar received the Doctor of Engineering Science degree (Eng.Sc.D.) from Columbia University, New York, N.Y. He is a registered professional engineer (P.E.) with the state of Kansas and a Member of the Institution of Electrical Engineers (British).

❖

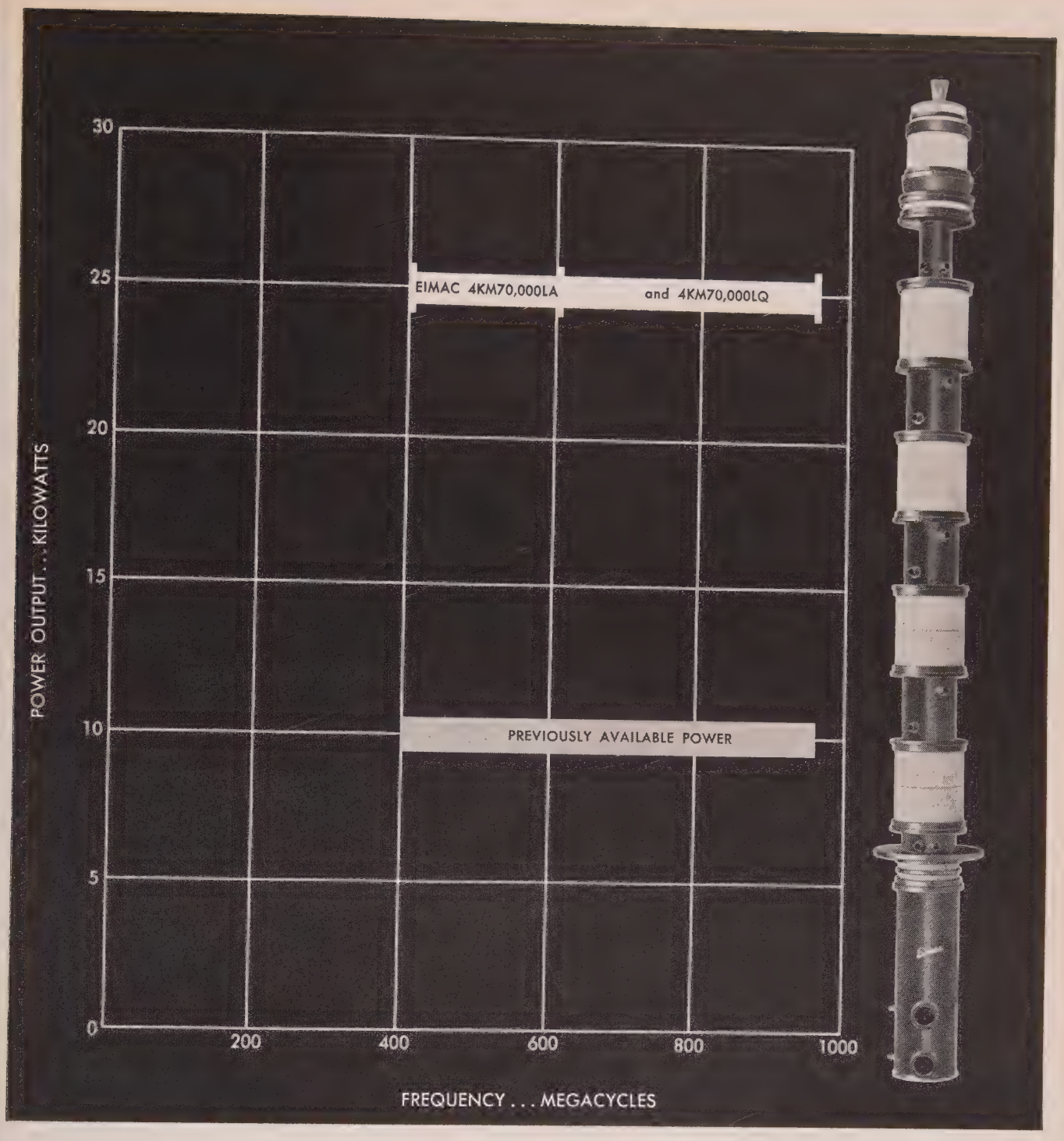
Earl J. Handly (A'47–M'55) has been promoted to the newly created post of division planning manager for Raytheon Company's Industrial Components Division. He has been market planning manager for the division since 1958. Mr. Handly will be responsible for coordinating the business planning activities of the division's four operations: industrial tube, industrial cathode-ray tube, receiving tube, and electromechanical components.

He joined Raytheon in 1944, and served in various engineering and manufacturing positions in the development of receiving and industrial radio tubes until 1954 when he was appointed product manager for gas and radiac tubes. He held this post until he was named market planning manager.

A graduate of the University of New Hampshire, he continued his studies at Boston University and Northeastern University. He is a member of the American Management Association, American Marketing Association, and is a licensed Professional Engineer in Massachusetts.

❖

(Continued on page 84A)



NOW...25KW AT 400-985 MEGACYCLES WITH EIMAC KLYSTRONS USING BeO RF WINDOWS

Eimac fills an important gap in the power vs. frequency chart with the introduction of the 4KM70,000LA and the 4KM70,000LQ. These new power-amplifier klystrons deliver minimum CW output of 25 kilowatts with a narrow band gain of 50db at frequencies from 400 to 985 megacycles. That's improved performance and higher power in this range than previously available.

The reason: these new ceramic and metal tubes are manufactured with windows of beryllium oxide—the amazing insulating material recently introduced by Eimac for vacuum tube use. It breaks through the problem of providing greater power-output capabilities in high power microwave tubes such as these by dissipating ever larger amounts of heat in dielectrics used as rf windows.

The 4KM70,000LA is designed for use at frequencies between 400 and 610 megacycles and the 4KM70,000LQ at frequencies between 610 and 985 megacycles. You'll find these tubes ideal for troposcatter communications and UHF television applications with an Eimac Klystron Amplifier Circuit Assembly to cover the specified frequency range. Eitel-McCullough, Inc., San Carlos, California.

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The Model STV is an extremely accurate and stable reference source for use with "null balance" devices such as potentiometers and other infinite impedance comparators. It is at least equivalent in accuracy to the best unsaturated standard cells and is superior in almost all other respects to both saturated and unsaturated types.

While the Model STV is essentially a zero current drain source, it can be operated into any impedance without damage. It can be short circuited indefinitely without affecting accuracy or life expectancy and it will almost instantaneously regain its original open circuit voltage when the short is removed. Vibration from transportation, exposure to extremes of temperature, and operating position do not affect its accuracy.

Specifications — Type "A"

Input: 90-135 v.; 60 cps; 25 va.

Output: 1.0000 v. and 1.0185 v.

Accuracy: ± .01% of nominal listed output (certificate furnished to .001% of actual output).

Stability: ± .005% of actual output, for 100-125 v. input and 20° – 30°C.; .01% for 90-140 v. input and 15° – 35°C.

Temp. Range: 15°C – 35°C (operates with reduced accuracy beyond these limits, but with its voltage exactly reproducible).

Operational Life: 25,000 hours minimum.

Size: 9⅜" x 7⅝" x 5". **Weight:** 10 lbs.

The Model STV is available for 19" rack panel mounting and in 3½" x 3" x 3" cans for OEM users (input must be regulated to 1%). Write for additional information on all types and special versions.

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ELECTRICAL INSTRUMENTS OF PRECISION SINCE 1927

IRE People

(Continued from page 82A)

Oliver Read (A'43), formerly publisher of *Electronics World*, editor and publisher of *Popular Electronics*, and publisher of *Hi Fi-Stereo Review* has joined the staff of the Instruction Department of the National Radio Institute, Washington, D. C.

The National Radio Institute, established in 1914 and considered one of the oldest and the largest home-study schools teaching only subjects related to electronics, television, radio, and electrical appliances, is accredited by the Accrediting Commission of the National Home Study Council, officially designated by the U. S. Office of Education as the accrediting agency in this field.

As Director of Publications, Mr. Read will be responsible for the NRI program of continuous revision of instruction materials, and will assist in planning and development of new home-study courses.

❖

Herbert H. Rosen (S'51–A'54–M'57) has been appointed Corporate Director of Public Relations at Hoffman Electronics Corporation, Los Angeles, Calif. Mr. Rosen recently resigned as Assistant Director for Educational Programs for the National Aeronautics and Space Administration.

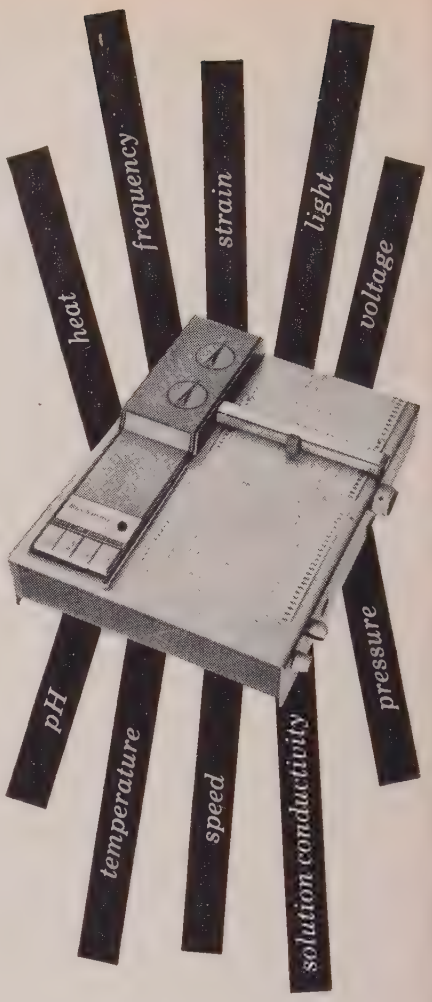
Before joining NASA in 1958 as the Deputy Director of the Office of Public Information, he was Washington Editor for Hayden Publications, reporting extensively on Washington affairs related to the electronics industry in *Electronic Week* and *Electronic Design*. He had previously been Chief of Technical Information for the ACF Electronics and Nuclear Energy Products Divisions of ACF Industries, Inc. and a member of the Office of Technical Information of the National Bureau of Standards, where he had started as a microwave standards engineer. Mr. Rosen was also a writer-announcer for several radio stations in New York and Louisiana. During World War II he served with the U. S. Marine Corps as a radar and gun data director technician.

Born in Philadelphia, Pa., Mr. Rosen attended the University of Pennsylvania, Moore School of Engineering and received the Bachelor of Science in Engineering degree from The George Washington University, where he also participated in the graduate program in engineering administration. He has served as President of the GW Engineer Alumni Association and as a vice president of the Governing Board of the General Alumni Association of that university.

Mr. Rosen is a member of the National Press Club, the Aviation/Space Writers Association, American Ordnance Association, the Armed Forces Communications-Electronics Association, and Sigma Tau. During his tenure in Washington, D. C., he served as a public relations advisor to AFCEA, IRE, and several local engineering organizations.

❖

(Continued on page 87A)



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The new Beckman Potentiometric Strip-Chart Recorder is low-cost and simple to operate. Compact, push-button unit features *continuously adjustable* input voltage span of 10 to 100mv full scale, facilitating its use with pH meters, spectrophotometers, and many other laboratory or process control instruments. Pen response is 1.0 second full-scale, and pen zero can be set to any point throughout the 5" pen travel. Chart speed is 1" per minute, and additional speeds are available with plug-in units. Other accessories include a circular chart recording adaptor and an external circuit controller. Unit may be wall-mounted or placed on laboratory bench. Ask your Beckman laboratory apparatus dealer for complete information, or write for Data File 9-11-21.

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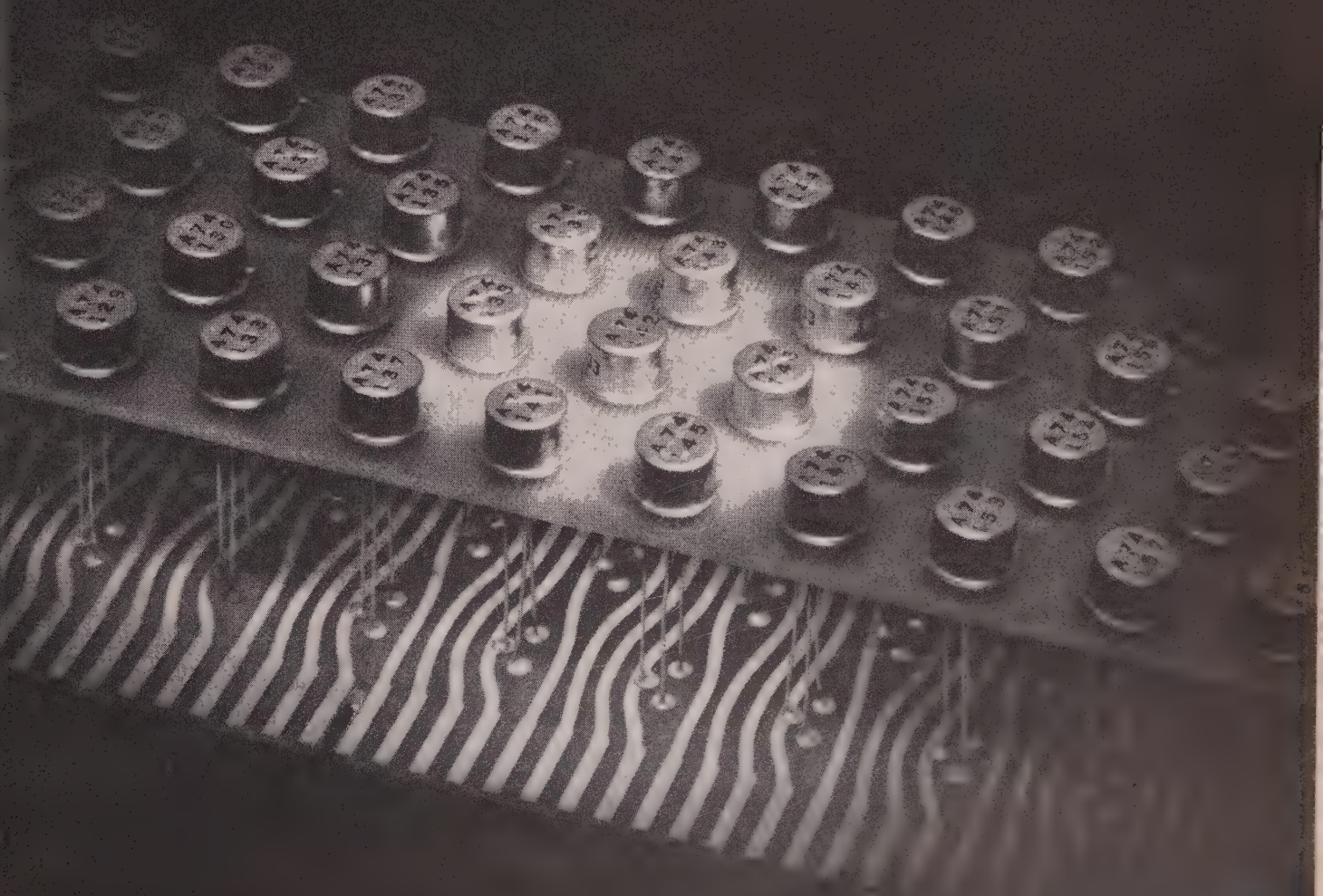

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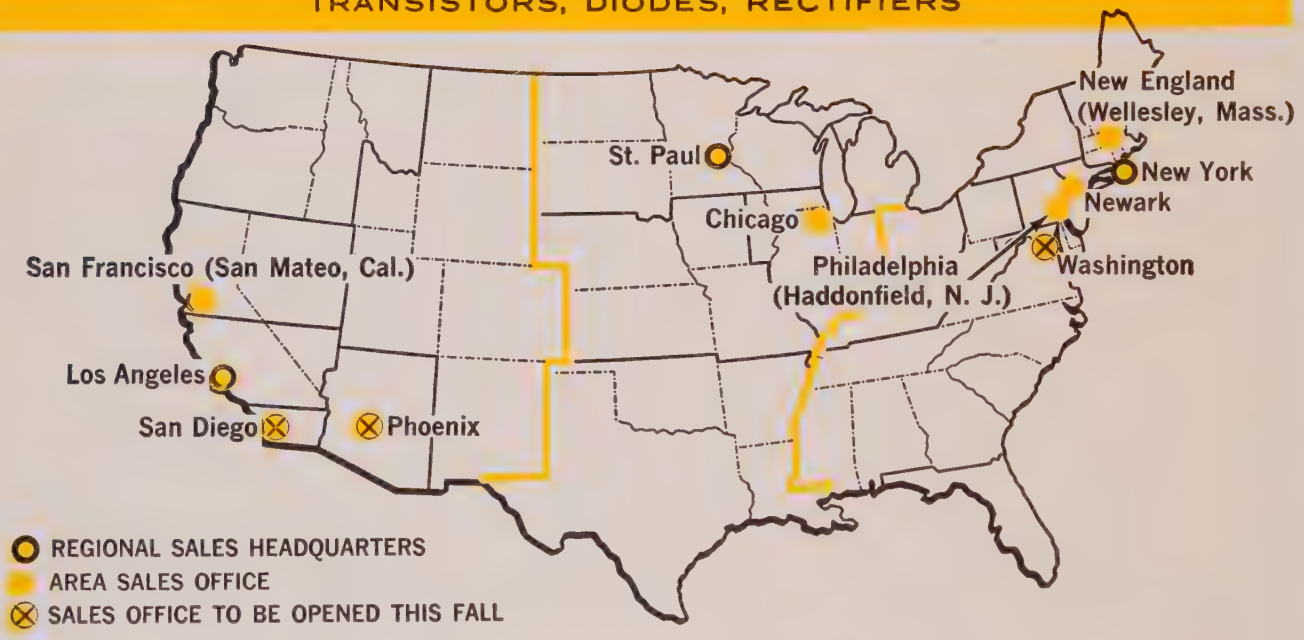

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IRE People

(Continued from page 84A)

Charles W. Skelton (SM'56) has been elected President of the newly organized International Data Systems, Inc., Dallas, Texas.

He holds four patents in the digital field, and he was the designer of the first digital programmer to be used in a ballistic missile. He was recently head of the Computer Branch in the Research and Development Department at Texas Instruments.

Mr. Skelton holds the degree in electrical engineering from Texas A & M College, College Station, and has attended Southern Methodist University, Dallas, Texas, Graduate School.

❖

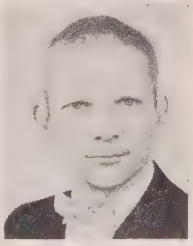
Joe H. Smith (S'47–M'53) has been elected Vice-President of the newly organized International Data Systems, Inc., Dallas, Texas.

He was recently head of the Instrumentation Section in Texas Instruments' Missile Department. He designed the first transistorized telemetry system used in a missile, and since then he has been responsible for the development and manufacture of all telemetry and instrumentation systems at Texas Instruments.

Mr. Smith holds the degree in electrical engineering from Purdue University, Lafayette, Ind.

❖

Herbert F. Somermeyer (A'53–M'58), has been appointed acting manager of the patent department at the St. Paul division of Remington Rand Univac, it has been announced.



H. F. SOMERMEYER

A registered patent attorney and member of the Minnesota Bar, he was employed by Univac as a logical designer and patent engineer prior to his association with the Minneapolis patent law firm of Carlsen and Carlsen.

Mr. Somermeyer received the electrical engineering degree from Iowa State University, Ames, in 1950. He graduated cum laude from the William Mitchell College of Law, St. Paul, Minn. in 1957.

❖

Appointment of **Edwin A. Speakman** (A'43–M'44–SM'48–F'45) as Division Vice President, Missile Range Programs, for the Government Services Department, RCA Service Company, has been announced.

In his new capacity, he will be in charge of RCA's Test Missile Project at Patrick Air Force Base, Fla., and other associated activities.

He joined RCA in 1958 as Manager of Planning for Defense Electronic Products and later was appointed Administrator for

(Continued on page 88A)

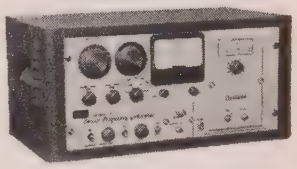
JERROLD

R. F. Test Equipment

Quantitative Measurements Using Sweep Frequency Techniques

Model 900A—THE MOST VERSATILE SWEEP GENERATOR $1,260.00

CENTER FREQUENCY—VHF 0.5 to 400 MC UHF 275 to 1000 MCS—SWEEP WIDTH—up to 400 MCS—FLATNESS—±0.5 db over widest sweep!



Model 707—ULTRA FLAT SWEEP GENERATOR $795.00

Featuring ±5/100 db flatness—Plug-in osc. heads*; variable sweep rates from 1/min. to 60/sec.; all electronic sweep fundamental frequencies; sweep width min. of 1% to 120% of C.F.

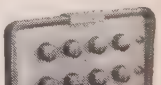
*Heads available within the spectrum 2 to 265 MCS

Models 601/602—PORTABLE GENERAL PURPOSE $295.00

COVERAGE—Model 601—12 to 220 MCS. Model 602—4 to 112 MCS—FLATNESS — ±0.5 db OUTPUT—up to 2.5 V RMS WIDTH-1% to 120% of C.F.

Model FD-30 $250.00

High speed DPDT coaxial switch permitting oscilloscope measurements without calibration—all measurements referenced continuously against standard attenuators.



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IRE People

(Continued from page 87A)

Electronic Intelligence Systems. Previously, he had been Vice President of the Fairchild Engine and Airplane Corporation and for five years was General Manager of Fairchild's Guided Missiles Division at Wyandanch, L. I., N. Y.

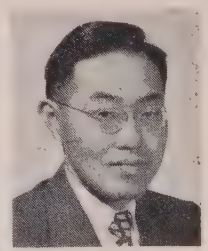
He was graduated from Haverford College, Haverford, Pa., with a degree in physics. He remained there for three years as an instructor in physics and as part of his graduate study invented a photoelectric timing system. In 1934, he joined the research laboratories of the Philco Corporation where he developed the telescopic rod antenna, now used on most radio-equipped automobiles.

In 1940, he was appointed Assistant Superintendent of the Radio Division, Naval Research Laboratory in Washington, D. C., and in 1945 became head of the Laboratory's Electronic Counter-measures Branch. During this period, he initiated developments in Naval radar equipment for which he received the Navy's Meritorious Civilian Service Award. In 1949, he joined the Research and Development Board of the Department of Defense as Executive Director of Electronics, and in 1950 he was made Vice-Chairman of the Board with responsibility to coordinate military research in electronics, guided missiles, aeronautics and atomic energy.

Mr. Speakman is an Associate Fellow of the Institute of Aeronautical Sciences, and has served for a number of years as a Consultant to the Department of Defense. He has served as Chairman of the Guided Missile Committee of the Aerospace Industries Association and as National Chairman of the Professional Group on Military Electronics.

❖

Dr. Kiyo Tomiyasu (S'41–A'42–M'49–SM'52) has transferred to the General Engineering Laboratory, General Electric Co., Schenectady, N. Y., as Microwave Engineer, RF and Communication Engineering, it was announced recently.



K. Tomiyasu

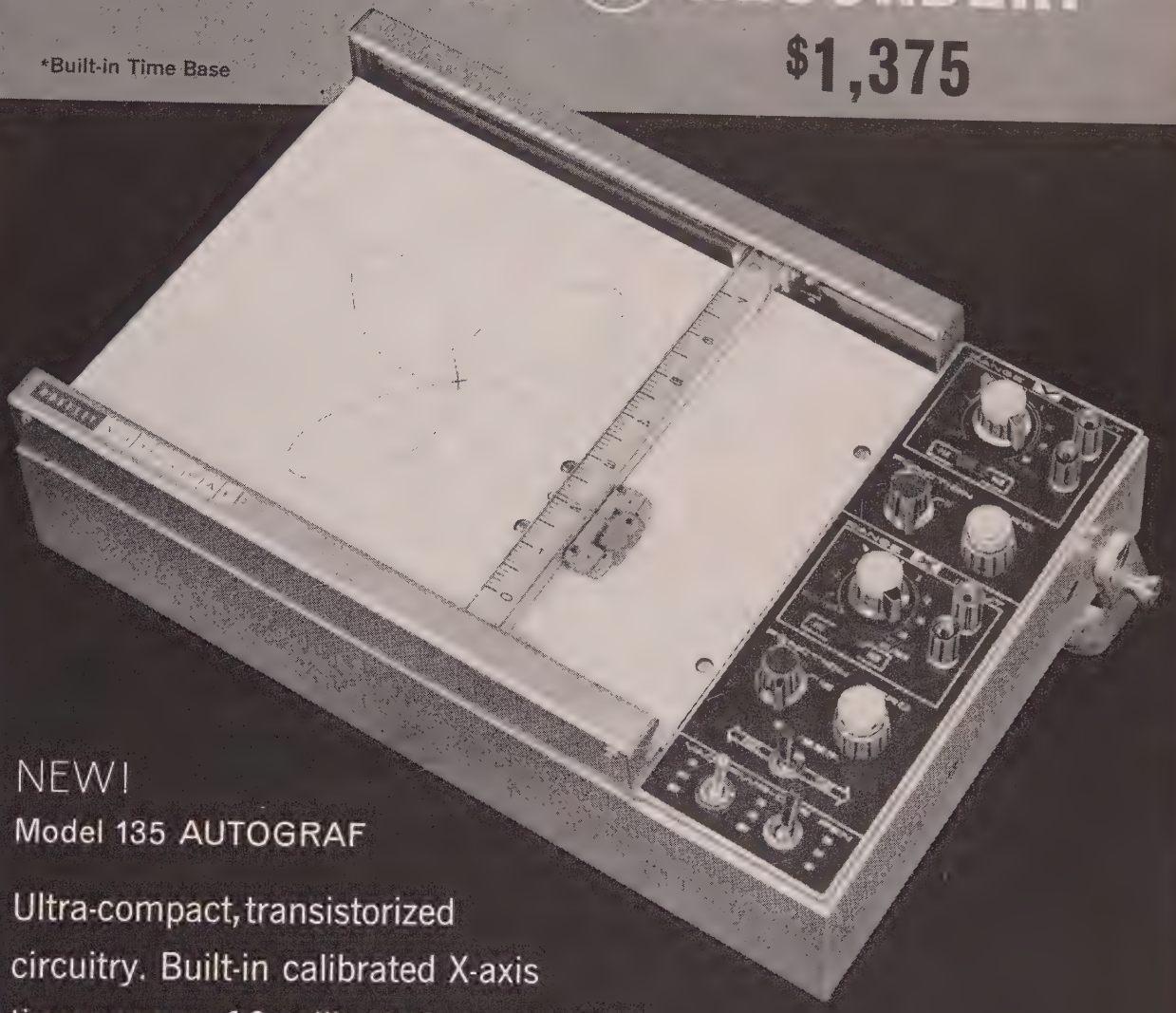
A native of Las Vegas, Nev., he received the B.S. (E.E.) degree from California Institute of Technology, Pasadena, in 1940 and the M.S. (E.E.) from Columbia University, New York, N. Y., in 1941. He also studied at Stanford University, Stanford, Calif., on a Low Scholarship and at Harvard University, Cambridge, Mass., as a Gordon McKay Scholar. He received the Ph.D. degree in Communication Engineering from Harvard in 1948, and served one year there as an instructor in the Lyman Laboratory.

From 1949 to 1955, he was with the Sperry Gyroscope Co., in charge of micro-

(Continued on page 92A)

World's Most Compact, Lightweight 8½" x 11" XY-(T)* RECORDER: $1,375

*Built-in Time Base



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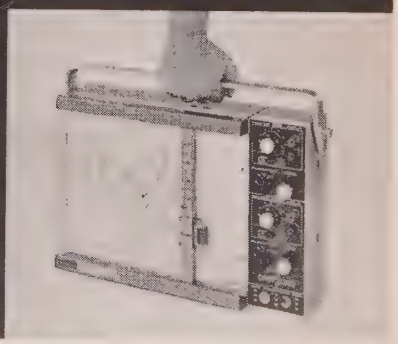

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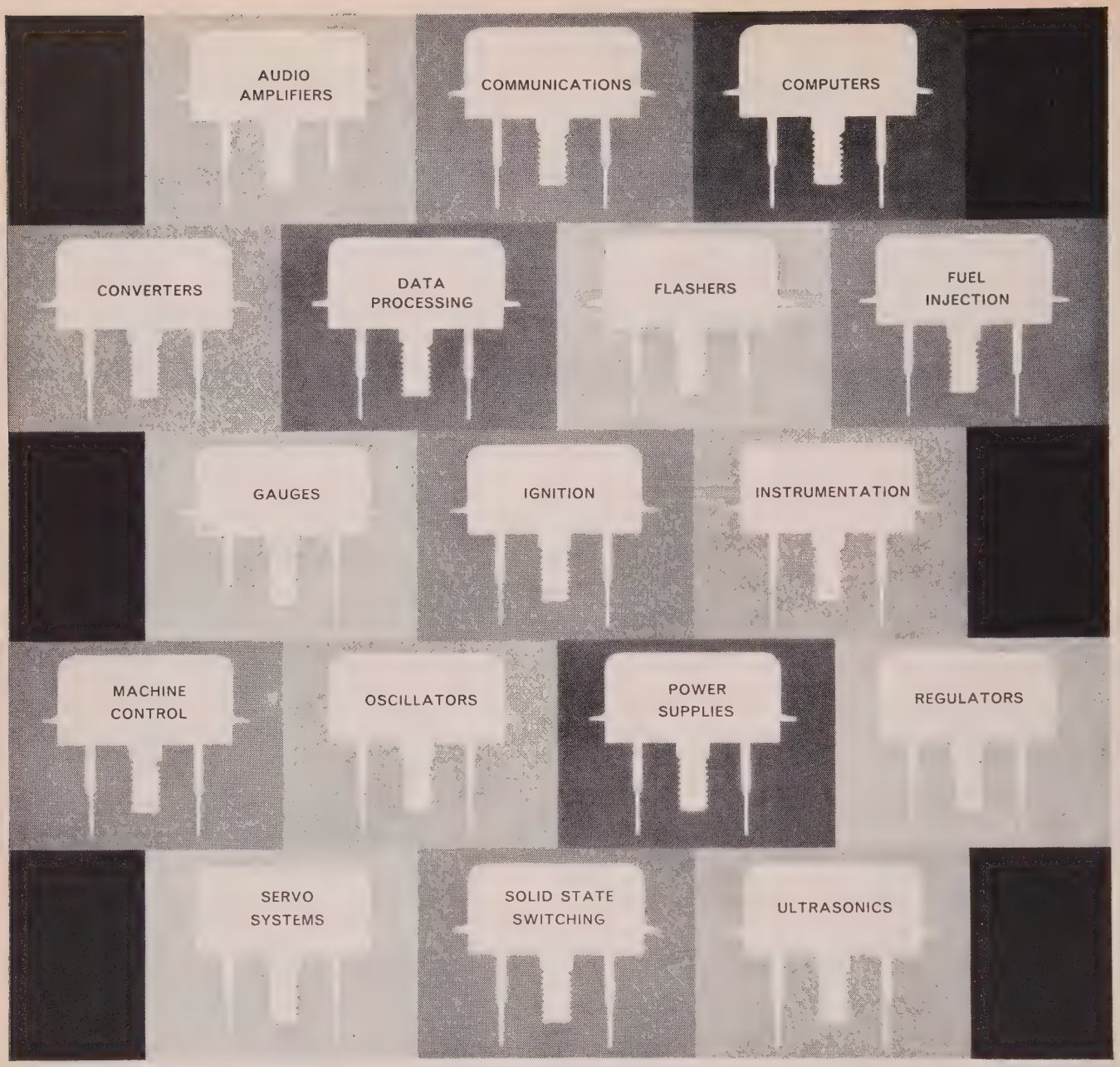

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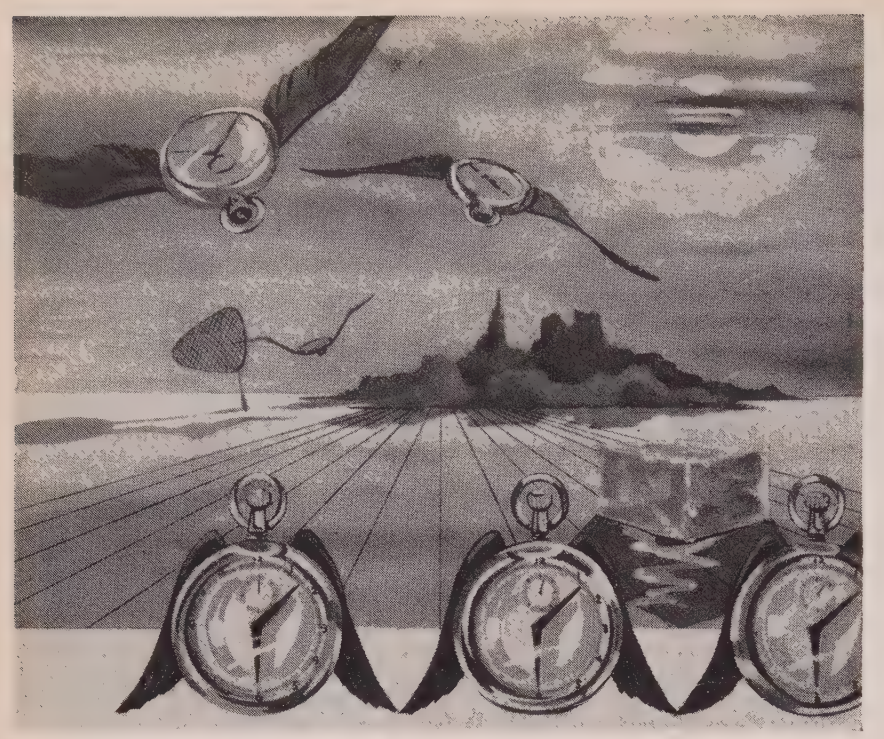

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IRE People

(Continued from page 88A)

wave research engineering. He then joined the General Electric Co. as Consulting Engineer in the Microwave Laboratory at Palo Alto, Calif. There he worked on projects related to microwave frequency memory, high power duplexers, filters, harmonic measurements in waveguide, electron ballistics, and numerous other problems. Since 1959, he has been consulting engineer to the High Power Klystron Sub-Section. He was Editor of the IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES for two years (1957–1958) and is presently a member of the Editorial Board.

Dr. Tomiyasu holds twelve patents and has authored 16 published papers. He is presently Chairman of the Administrative Committee of the Professional Group on Microwave Theory and Techniques. He is also a member of the American Physical Society and of Sigma Xi.

❖

Dennis D. Willard (S'50–A'51–SM'58) IBM executive credited with numerous engineering innovations, has been named Manager of the Advanced Computer Laboratory for Ampex Data Products Company, it was announced recently.

D. D. WILLARD

Formerly Manager of Advanced Technical Planning, Advanced Systems Development Division at IBM, San Jose, Calif., he joined Ampex in July, 1960 to manage the Advanced Computer Laboratory, encompassing both Advanced Development and Systems work for the Computer Products Division.

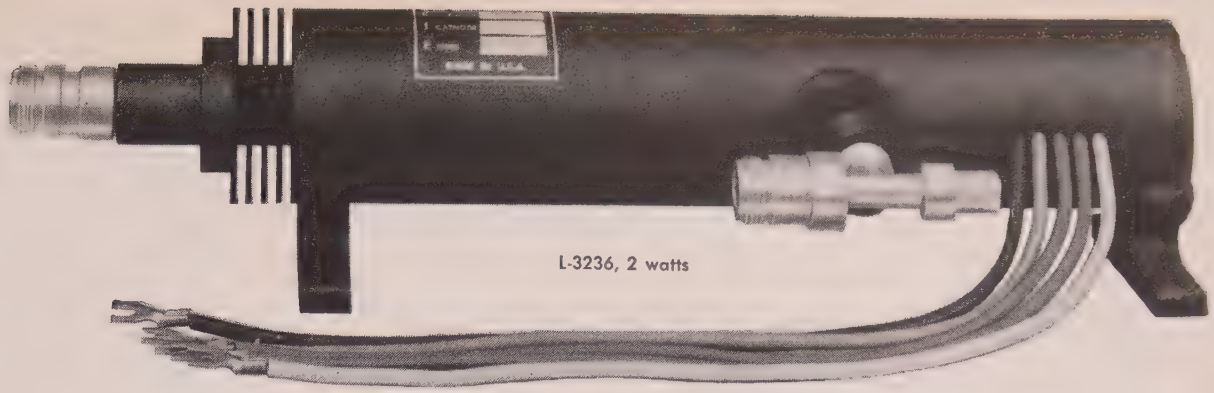
He received the B.S. (E.E.) Degree from the University of Kansas, Lawrence, graduating with honors in 1950. Joining IBM at Endicott, N. Y. that same year, he served in a variety of assignments in design, release, and advanced development in both the commercial and military divisions prior to his transfer to San Jose in early 1956.

Mr. Willard is a member of the Professional Groups on Electronic Computers and Industrial Electronics and also holds membership in Tau Beta Pi and the Society for the Advancement of Management. He has been a committee member of both the 1959 and 1960 Western Joint Computer Conferences, serving as Chairman of the 1960 WJCC Publications Committee.

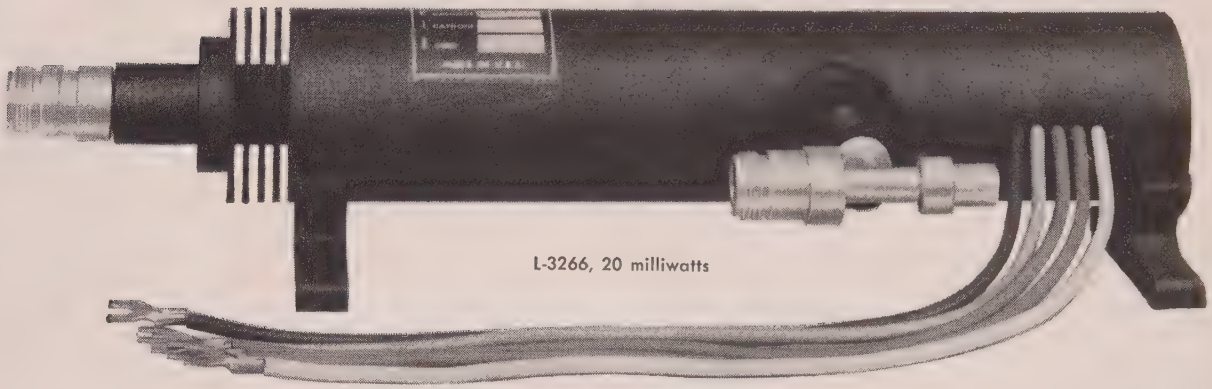
❖

Harry E. Stockman (A'42–M'44–SM'45) has joined the faculty of Lowell Technological Institute, Lowell, Mass., as professor in the Division of Physics and Engineering Science.

(Continued on page 95A)



TWTwins

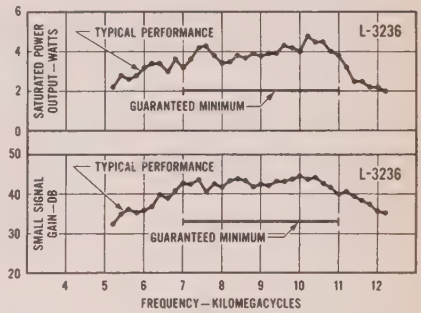


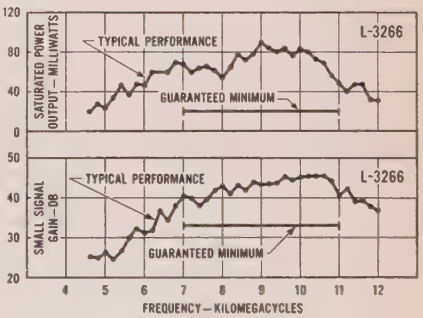
These Litton TWT twins are PPM focussed X-band traveling wave tubes. They are *not* prototypes. They are metal and ceramic tubes in field application *now*.

Designed to cover X-band with minimum saturated CW power of 20 milliwatts (L-3266) and 2 watts (L-3236), they may be operated in cascade to amplify signals as small as −50 dbm to the 2 watt level. Their performance, far exceeding the conservative specifications, is evident in the accompanying graphs.

Small size (less than 12″ in length), light weight (under 4 pounds each), and extreme environmental capability (temperature compensated −54°C to 86°C) make these tubes the ideal choice for military applications. A typical airborne equipment, designed and manufactured by Granger Associates of Palo Alto, California, incorporates the L-3266 and L-3236 and occupies only 0.75 cubic feet, including all necessary power supplies, modulating circuitry, cooling, etc. This equipment is now in field operation.

If your work involves ECM repeaters, radar target enhancement, frequency diversity radar or any application requiring broadband microwave amplifiers, appraise these new tubes. In production quantities their price is the lowest in the field. Ask for catalog sheets on the L-3266 and L-3236. Address: Litton Industries Electron Tube Division, 960 Industrial Road, San Carlos, California.





(see them at WESCON)

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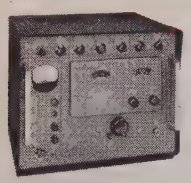
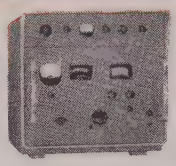
500 mc to 50,000 mc

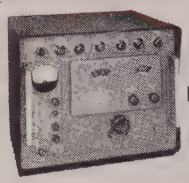
MOST FEATURES

MODEL MSG-I

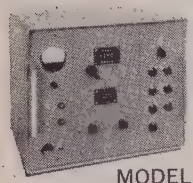


MODEL PMR

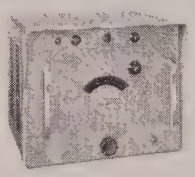




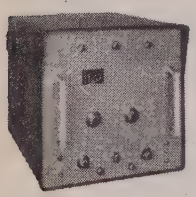
MODEL MSG-2A



MODEL KSS

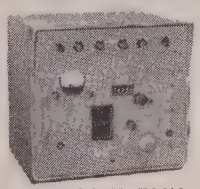


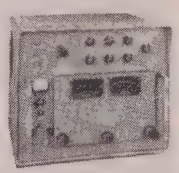
MODEL MSG-34



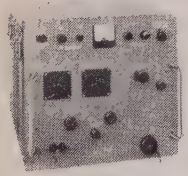
MODEL CSG

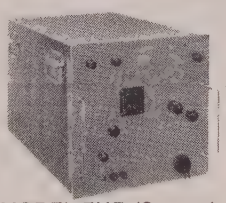
MODEL PMX





MODEL PMK





MODEL EHF (Generator)

MODEL EHF (Source)

MODEL PMR	500 to 1,000 mc	Complete modulation capabilities — internal pulse modulator or FM modulator
MODEL MSG-1	950 to 2,400 mc	Complete modulation capability including square wave modulation
MODEL MSG-2A	2,000 to 4,600 mc	Complete modulation capability including square wave modulation
MODEL MSG-34	4,200 to 11,000 mc	Widest frequency range in a single instrument
MODEL KSS	1,050 to 11,000 mc	Compact high power signal source with plug-in tuning units — internal modulation
MODEL PMX	4,450 to 11,000 mc	Calibrated 1 milliwatt signal generator with complete modulation capability
MODEL CSG	1,000 to 16,000 mc	Higher power sweep generator
MODEL PMK	10,000 to 21,000 mc	Wider modulation capabilities — calibrated 10 milliwatt output
MODEL EHF (generator)	18,000 to 39,700 mc	High frequency signal generator — operates on fundamentals
MODEL EHF (source)	18,000 to 50,000 mc	Widest and highest continuous frequency range — operates on fundamentals

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Please send me information and specifications on

- ☐ Model PMR
- ☐ Model MSG-1
- ☐ Model MSG-2A
- ☐ Model MSG-34
- ☐ Model KSS
- ☐ Model PMX
- ☐ Model CSG
- ☐ Model PMK
- ☐ Model EHF (generator)
- ☐ Model EHF (source)
- ☐ A copy of "Notes on Microwave Measurements."

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IRE People

(Continued from page 92A)

A native of Stockholm, Sweden, Dr. Stockman received the E.E. degree from Stockholm Technical Institute, the M.S. degree from the Royal Institute of Technology in Stockholm, and the Sc.D. degree from Harvard University. He previously taught at those three institutions, and at Merrimack College where he was professor and chairman of the Department of Electrical Engineering. Dr. Stockman also has been associated with Norden Ketay, Boston division, as research and development manager, and with the Air Force Cambridge Field Station Materiel Command as chief of the communications laboratory. He is a consultant to industry in the field of electronics.

❖

John T. Underhill (A'54–SM'57) has been appointed Operations Manager of Cubic Corporation, San Diego, Calif. Formerly Assistant Chief Engineer, he has also been named Assistant to Executive Vice-President, R. V. Werner, and as Operations Manager will be responsible for direct line operations management of the engineering department and for performance on all engineering projects.

He is an electrical engineering graduate of the University of Arizona, Tucson. A field test engineer, senior design engineer, staff member and project engineer with the Sandia Corporation, he was also employed by Convair before joining Cubic in 1957.

Mr. Underhill is a member of the American Institute of Electrical Engineers, the American Ordnance Association, Tau Beta Pi, Pi Mu Epsilon, Sigma Pi Sigma, and Eta Kappa Nu.

❖

Election of **W. Walter Watts** (A'53) as Director of the Radio Corporation of America has been announced.

He became Chairman of the Board and President of the RCA Sales Corporation on August 18, 1960 in addition to his post as RCA Group Executive Vice President.

He joined RCA in 1945 after wartime service as a Signal Corps Colonel in distribution and procurement. Long experienced in the sales and distribution phases of the electronics industry, he has held a broad range of executive positions of increasing responsibility in RCA.

He advanced from General Sales Manager of the Engineering Products Department of the former RCA Victor Division to Vice President of RCA in 1946, and in 1954 he was elected Executive Vice President, Electronic Products. The following year, the Electronic Components unit was placed under his direction. In 1958, he was elected Group Executive Vice-President, responsible for the RCA Electron Tube Division, the RCA Semicon-

(Continued on page 98A)

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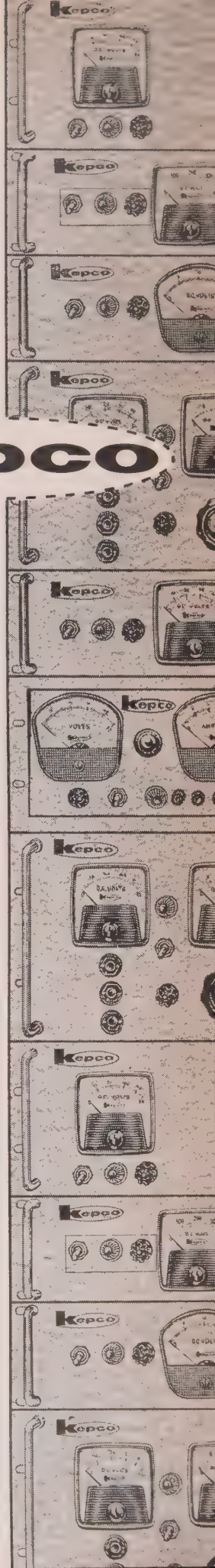

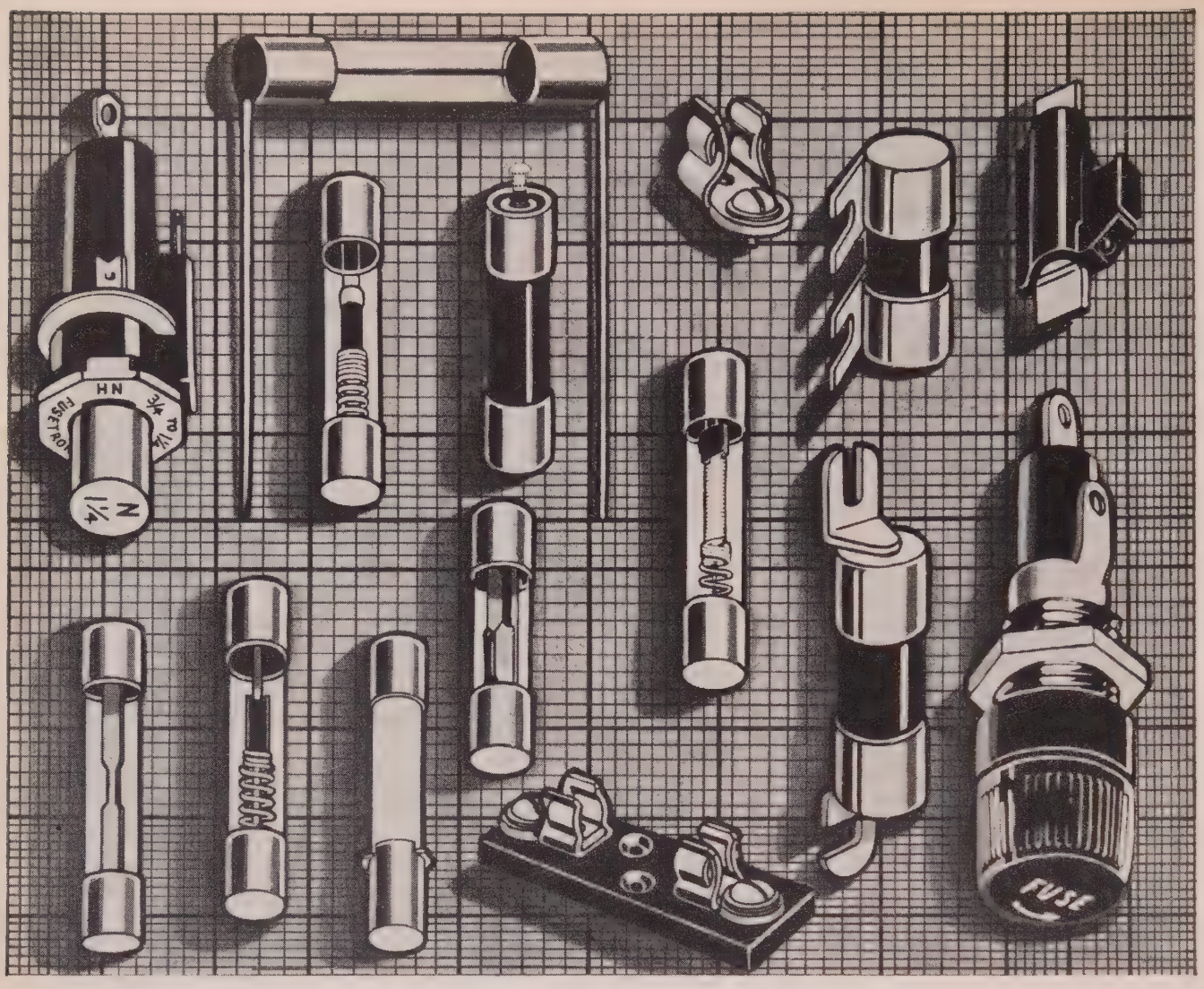

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You get double protection when you specify BUSS fuses.

First, BUSS fuses are designed and manufactured to give maximum electrical protection.

But it doesn't stop there.

Every BUSS fuse is carefully tested in a sensitive electronic device that automatically rejects any fuse not correctly calibrated, properly constructed and right in all physical dimensions.

This is your assurance that BUSS fuses will operate as intended.

Second, BUSS fuses help protect your reputation as a quality manufacturer. A fuse that opens prematurely causes a needless shutdown. Likewise, one that doesn't function properly may cause other components to burn out or be damaged. In either case, it's an annoying headache for the customer who buys your equipment. More often than not, he will blame your product for his trouble.

With dependable BUSS fuses, you need have no worries that faulty fuses will give your product a bad name. That's why it makes good sense to specify BUSS fuses.

For more information on BUSS and FUSETRON small dimension fuses and fuseholders . . . Write for bulletin SFB.

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BUSS FUSES ARE MADE TO PROTECT - NOT TO BLOW NEEDLESSLY

FUSETRON BUSS — TRUSTWORTHY NAMES IN ELECTRICAL PROTECTION — McGRAW EDISON

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Typical field application of NF-105 in mobile Radio Interference Laboratory

Assure Reliability under Adverse Conditions...

in the Field and in your Plant

SPECIFY EMPIRE'S COMPACT AND RUGGED

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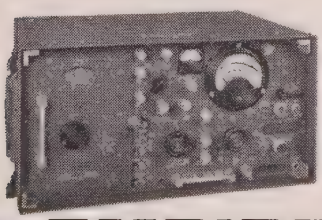
MODEL NF-105

- Measures 150 kilocycles to 1000 megacycles accurately and quickly with only one instrument.
- For measurements in accordance with Specifications: MIL-I-6181B, Class 1; MIL-S-10379A; MIL-I-11683B; MIL-I-11748B; MIL-I-12348A; MIL-I-13237; MIL-I-16910A; MIL-I-26600 (USAF), Category A; F.C.C. Specifications.
- Direct substitution measurements by means of broad-band impulse calibrator, without charts, assure repeatability.
- Economical...avoids duplication.
- True peak indication by direct meter reading or aural slideback.
- Four interchangeable plug-in tuning units, for extreme flexibility.
- Safeguards personnel...ALL antennas can be remotely located from the instrument without affecting performance.
- Self-calibrating, for reliability and speed of operation.
- Compact, built-in regulated "A" and "B" power supply, for stability.
- Minimum of maintenance required, proven by years of field experience.

DELIVERY FROM STOCK



The unique design of Model NF-105, with 4 plug-in tuning units, avoids costly repetition of circuitry and components common to all frequency ranges, at savings in size, weight and cost. Simple to operate, this instrument permits fast and accurate measurements of both broadband or CW signals. *Send for our Catalog*

Plan to attend our next seminar on interference instrumentation, details upon request.

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superb new NULL DETECTOR

The new Keithley 151, incorporating a unique photo-conductive modulator of Keithley design, is useful wherever a suspension galvanometer can be used, and where a galvanometer is not sufficiently sensitive, fast or rugged. Currents as low as 2×10^{-13} ampere can be detected.

Ranges: 11 linear ranges in 1x and 3x steps, from 100 μv to 10 v f.s.; 5 non-linear ranges, 0.001 to 10 v f.s., each covering three decades.

Accuracy: Linear ranges, $\pm 3\%$ of f.s.; non-linear, $\pm 10\%$ of input.

Input Resistance: 10 megohms on all ranges. Max. power sensitivity over 10^{-17} watt.

Response Speed: On 100 μv range, 2.5-sec.; 1-sec. on all others.

Noise: Below 2% f.s. all ranges.

Zero Drift: Less than 10 μv per day.

Output: 10 volts at 1 ma f.s.

Price: 151 Cabinet Model . $395.00
151R Rack Model . $385.00

KEITHLEY INSTRUMENTS

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CLEVELAND 6, OHIO

IRE People

(Continued from page 95A)

ductor and Materials Division and the RCA International Division.

Last July 1, he was assigned executive responsibility over the Home Instrument, RCA Victor Distributing Corporation, and Distributor and Commercial Relations activities of RCA.

Born in Chicago, Mr. Watts attended the Lane Technical School and Chicago Telegraph Institute there. In 1958, the United States Senate confirmed his appointment to the rank of Brigadier General, United States Army Reserve. He is past President of the Armed Forces Communications and Electronics Association, and a member of the American Society of Naval Engineers, the American Relay League and the American Ordnance Association.

Professional Group Meetings

Audio

Chicago—September 9

"Receiver Design Considerations for Sterophonic FM Multiplex Broadcasting," C. Eilers, Zenith Radio Co., Chicago, Ill.

Automatic Control

Chicago—September 9

"The Automatic Focusing System of the Human Eye," J. Warshawsky, Northwestern University, Evanston, Ill.

Los Angeles—May 10

"Adaptive Cross-Correlator," G. W. Anderson and R. Buland, Aeronutronics Corp., Newport Beach, Calif.

Bio-Medical Electronics

Chicago—September 9

"A Miniature Transistorized Cardiotachometer for Monitoring Ambulatory Subjects," L. J. Ryan, Ill. Bell Telephone Co., Chicago, Ill.

Broadcasting

Florida West Coast—August 31

"Parameters of Reliability," C. M. Ryerson, RCA, Camden, N. J.

Electron Devices

San Francisco—September 8

"New Ideas on Thermionic Emission," E. A. Coomes, Univ. of Notre Dame, South Bend, Ind.

(Continued on page 104A)

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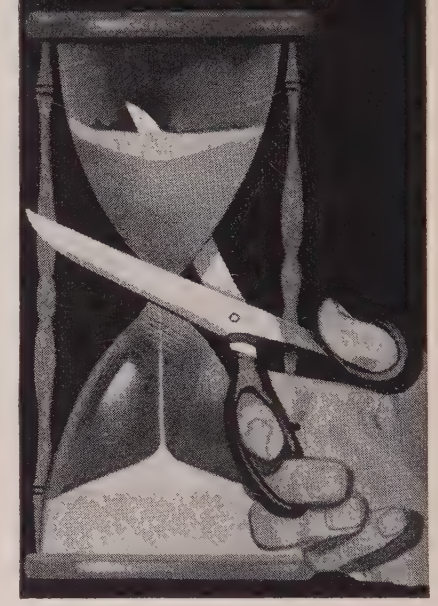
SPECIFYING DEFLECTION YOKES



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CONTROLS COMPANY of AMERICA develops first solid-state moisture-sensing system.

New magnitude of reliability introduced to moisture control field

Here's a new system that precisely measures moisture content—with proven reliability. In fabrics . . . organic or inorganic compounds . . . mixes — in fact wherever moisture is a problem this system offers appliance and industrial engineers a new standard of accurate moisture control.

The control continuously measures moisture content with an instantaneous control response and can be used for high volume applications or single batch units. Controls Company moisture-sensing systems are already in production.

In designing their moisture-sensor, quality conscious engineers at Controls Company of America in Schiller Park, Ill. specified TI transistors with linear beta, high power gain and low distortion characteristics.

Quality TI transistors for your particular application are available in production quantities through your TI distributor or TI sales office.

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Glass-enclosed, fusion-sealed Corning NF resistors have boiled merrily in salt water for days without showing a jot of change in their electrical characteristics.

These are resistors that are rugged, completely moisture resistant, highly vibration resistant . . . in short, resistors that exceed the requirements of MIL-R-10509C, Char. B, better than any we've seen or heard of.

The key to such fortitude is our NF structure. We start with glass rods with metal oxide applied under heat. This in itself makes a moisture-resistant, almost abnormally stable resistor, as you well know if you have ever used our regular N-style resistors.

We encapsulate this basic unit in a glass envelope and apply glass-to-metal seals at the leads . . . comparable to those in a vacuum tube.

IMMEDIATE DELIVERY • There are two models of this gem in production, ready for *quick shipment:* the ⅛-watt NF-60 and the ¼-watt NF-65. Resistance ranges from 100 ohms to 360K ohms. Voltage ratings are 250v and 300v. Full rating at 70°C. with derating to 150°C. More data:

Load life	0.3%
Voltage coefficient	0.001%/v
Temp. coefficient	0.03%/°C.
Insulation resist.	100,000 megohms

To get this and other data for your file, just write and ask for Data Sheet CE-2.02. Address: Corning Glass Works, 542 High Street, Bradford, Pennsylvania.

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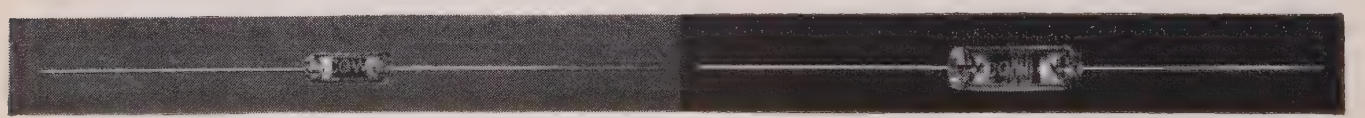

⅛-WATT NF-60 ACTUAL SIZE ¼-WATT NF-65

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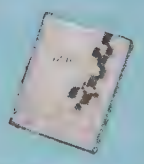
Life — over 700 hours reported
Peak power available — more than 10 kw. (More power than you can get from any other device at this frequency)
Duty cycle — up to 0.001. (For the BL-221, it is 0.00055)
Vibration — will survive 10 g's
Shock — 50 g's at 4 millisec
Lightweight — 7.25 lbs
Mounting — mates to modified standard flange
Ruggedized — Ceramic and metal construction
Fixed tuned

Band	Tube Type	Frequency Range (MC)	Minimum Peak Power (KW)	Output Mates with
V	BL-235	51,000–54,000	10	UG385/U
V	BL-236	54,000–57,000	10	UG385/U
V	BL-237	57,000–60,000	10	UG385/U
V	BL-221	69,000–70,500	10	UG385/U

New product catalog available. Send for your copy today.



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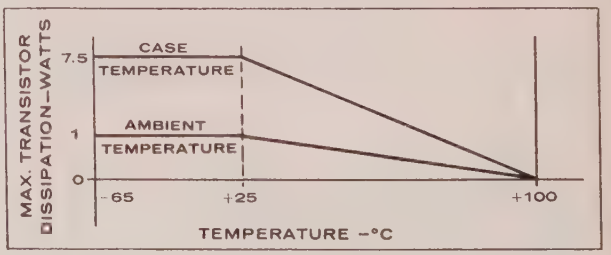
2N 1183	−35v	−20v	−45v	−20v	20-60
2N 1183A	−50v	−30v	−60v	−20v	20-60
2N 1183B	−60v	−40v	−80v	−20v	20-60
2N 1184	−35v	−20v	−45v	−20v	40-120
2N 1184A	−50v	−30v	−60v	−20v	40-120
2N 1184B	−60v	−40v	−80v	−20v	40-120



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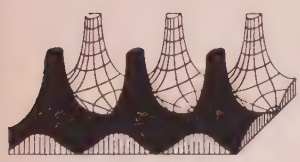
Poles and Zeros



September 22, 1960
Paris, France

Professor Ferdinand Hamburger, Jr.
Editor, The Institute of Radio Engineers
1 East 79 Street
New York 21, New York

Dear Ferd:

Here is a progress report from Gay Paree, a very important electronics center. To date, I have spent ten days at the General Assembly of the International Scientific Radio Union (URSI) in London, made a side trip to Edinburgh (visiting Professor Farvis of the University of Edinburgh and Mr. Stewart of Ferranti), and participated in the International Symposium on Data Transmission at Delft, Netherlands. The Data Transmission Symposium is co-sponsored by the IRE Benelux Section, the Koninklijke Institute van Ingenieurs (Royal Institute of Engineers), and the Nederlands Radiogenootschap (Netherlands Radio Society). The organizing committee, including Messrs. Rinia and Barrow of the Benelux Section, produced an excellent program. Attendance was over 500, and included engineers and scientists from all over the world—20 from the United States. Especially noticeable was the excellent spirit of cooperation between the IRE Benelux Section and the two Netherlands Societies. The conference was really an important step in IRE's increasing activity in international electronic affairs.

IRE, incidentally, was very well represented at the URSI General Assembly by Director Lloyd V. Berkner, President of URSI, 1959 IRE President Dr. Ernst Weber, 1957 IRE President Dr. John Henderson, and others. IRE Vice President J. A. Ratcliffe and his fellow United Kingdom committee chairmen efficiently organized and carried out the many aspects of the General Assembly program. In addition to the high quality of the various commission sessions the most noticeable feature of the conference was the hospitality of our British hosts. It can be described in one word: Superb.

But why make this trip? You remember, I am sure, the action of the IRE Board of Directors at its October, 1959, meeting in Toronto, when the following resolution was passed and put into effect. "It is understood that to promote the objectives of IRE generally, the IRE welcomes the acquisition of members in all countries and offers its services in augmenting their activities in the promotion of radio and electronics as a profession, with the further understanding that the IRE does not seek to displace any similar activities which might be taken on a national basis."

Since one of the pleasant duties of the IRE President is to visit as many IRE Sections each year as possible it is especially desirable that IRE Sections outside North America be included in the itinerary where possible. Incidentally, the distance from Boston to Paris is about half that from Boston to Honolulu. But, and much more important, so far as I know an IRE President has not heretofore officially visited the European IRE Sections, and today, the European electronic centers are making very important contributions to electronic progress.

Now, what about the rest of the trip? In Paris, I have just met with the officers of the Société Française des Electroniciens et des Radioélectriciens, and I also addressed a formal meeting of the Société. An official visit, including a talk to a meeting of the Section, is scheduled for the Milan Section of IRE. Informal meetings are also arranged in centers where there are appreciable numbers of IRE members, but where no Sections exist as yet. These centers include Geneva, Zurich, Copenhagen and Stockholm. In all, there are today about 1200 IRE members in the countries which are being visited.

IRE already is active in international electronic affairs, and enjoys friendly relations with sister societies in many countries. The basic purpose of this trip is to stimulate further IRE international effort, and to cultivate closer cooperative relations with professional radio and electronic societies in other countries.

With best personal regards,

Sincerely,

R. L. McFarlan

Ronald L. McFarlan
President

J. B. Russell

Director, 1960–1961

Dr. John B. Russell (A'36–SM'45–F'58) was born in Boston, Mass. on March 10, 1907. He received the B.S. degree in 1928, the M.S. degree in 1931, and the D.Sc. degree in 1950 from the Massachusetts Institute of Technology, Cambridge, all in the field of electrical engineering. He was an instructor at M.I.T. from 1929 to 1931, and then joined the Electrical Engineering Department of Columbia University, New York, N. Y., as an instructor. He was made an Assistant Professor in 1935, an Associate Professor in 1946, and a full Professor in 1949. In 1952, he became head of the Department of Electrical Engineering which he directed until joining the General Electric Company in 1955. In addition to his teaching and administrative duties, he was a member of Columbia's Electronics Research Laboratory which he helped organize in 1951. In 1948, he spent a seven-month sabbatical leave as a research associate at M.I.T.'s Research Laboratory of Electronics working in the field of oscillations in nonlinear systems.

For five years, during and shortly after World War II, he was on leave from Columbia University to the Government. From 1941 to early 1945, he was a Technical Aide to Division 7 of the National Defense Research Committee, organizing and directing research and development here and in England in the field of airborne fire control. He was Chairman of the Army-Navy, NDRC Airborne Fire Control Committee, and was Consultant to the three armed services on this subject. As a result of this work, he was appointed Expert Consultant in the Office of the Secretary of War in April of 1945. In this position, he acted as an advisor on the organization of a quantitative testing facility for the Air Force Proving Ground. During the summer of 1946, he was Expert Consultant in the Office of the Chief of Staff of the War Department and was concerned with the initial organization of the Research and Development Division of that office. For this work, he received the Army-Navy Certificate of Appreciation, and in 1946 the Naval Ordance Development Award with the Certificate of Exceptional Service.

Dr. Russell joined General Electric as a staff member of the Electronics Laboratory in 1954, while on leave from Columbia University. He was initially concerned with the application of special magnetic materials. He was made Supervisor of the Video Display Unit in March, 1955; Consulting Engineer for Magnetics and Dielectrics in July, 1955, having resigned from Columbia University; and Manager of Engineering Analysis in November of that same year. On January 1, 1957, he was appointed Manager of the Electronics Laboratory with the responsibility of directing the Company's principal activities in applied research and development in the broad field of electronics.

Throughout his academic career, he acted as a consultant on electronics to a number of industrial firms and government agencies such as Racon Electric Company, Radio Receptor Company, Hazeltine Research, IBM, the Franklin Institute, the Central Intelligence Agency, and the Air Research and Development Command.

Dr. Russell is a member of Epsilon Chi, Tau Beta Pi, and Sigma Xi. He is also a Fellow of the New York Academy of Sciences and the AIEE. He has served as a member and as Chairman of a number of technical committees of both the IRE and the AIEE. He is also a member of the American Society for Engineering Education, and the Association for Computing Machinery, which he helped organize.

Scanning the Issue

Interrelation and Combination of Various Types of Modulation (Meewezen, p. 1824)—It is not the policy of the PROCEEDINGS to reprint papers from other journals. Once in a while, however, a paper comes along which the reviewers and editors feel will be of wide and timely interest to the membership, and an exception is made. This paper is one such exception. It originally appeared in the *Proceedings* of our sister society, the Institution of Radio Engineers in Australia. It discusses and compares six types of modulation: amplitude modulation, suppressed carrier double sideband, single sideband, quadrature modulation, phase modulation and frequency modulation. In addition to being clearly written and easy to read, the paper has the virtue of making very clear the difference between quadrature and phase modulation and of pointing out why both types of modulation deserve more attention than they have received in the past. The discussion of combined forms of modulation is especially timely since it analyzes two of the AM stereo broadcast systems which have been proposed to the National Stereophonic Radio Committee of the EIA.

A Categorization of the Solid-State Device Aspects of Microsystems Electronics (Lesk, *et al.*, p. 1833)—Because of the many recent developments in the solid-state field, we have entered the world of the "micro." Today we find our language generously (and indiscriminately) sprinkled with terms like microminiaturization, microelectronics, microcircuits, micromodules and microsystems, not to mention molecular electronics, functional blocks and integrated devices. It is not surprising that there is considerable confusion as to what some of these terms mean and how they relate to one another and to specific solid-state devices. The authors of this paper attempt to restore some semantic order to this Babel-like scene by proposing four categories into which all solid-state devices may be placed; namely, basic, complex, integrated, and array. An excellent bibliography is included and the examples of new "complex" and "integrated" devices are in themselves of considerable interest.

A Two-Color Input, Two-Color Output Image Intensifier Panel (Nicoll and Sussman, p. 1842)—The PROCEEDINGS has reported a number of important developments in the photoconductive-electroluminescent device field during the past five years, to the point where it has become a primary source of information in this young field. This is the first time, however, that a multicolor device of this type has been described in the PROCEEDINGS, or, for that matter, anywhere else. The multicolor feature has been made possible by the recent advent of photoconductive powders with different spectral responses and electroluminescent phosphors of more than one color. The result is a novel two-color image intensifier panel in which one color is excited by infrared and the other by visible input images.

Construction of a Thermionic Energy Converter (Block, *et al.*, p. 1846)—Great strides have been made in the last half-dozen years in developing new methods of converting energy from one form to another, leading to important new sources of electrical power such as the solar cell. Today, in an age of plasmas, nuclear fuels and rocket propulsion, there is particular interest in the direct conversion of heat to electricity. In addition to thermoelectric converters and magnetohydrodynamic generators, one of the promising approaches under investigation is the thermionic converter, a vacuum or gas-filled device containing a hot cathode and a cold cathode. This paper contains valuable information on the design, construction, and testing of a thermionic converter of the cesium vapor type. The topic will be of interest not only to energy conversion specialists but also to those concerned with new power sources, especially those which can run unattended for long periods of time.

Nanosecond Response and Attenuation Characteristics of a Superconductive Coaxial Line (Nahman and Gooch, p. 1852)—Several important developments have been reported recently involving the use of superconductive elements. In each instance the main point of interest was not so much in the superconductor per se, but rather in the fact that the element could be switched back and forth from a superconducting to a normal state. The development reported here considers the seemingly obvious, but nonetheless novel, application of a superconductor simply as a "super" conductor. A superconductive coaxial line is described whose specific use is to act as a low distortion delay line for nanosecond pulse oscillography. However, the results are of broad import to the general field of nanosecond pulse transmission. In particular, it is shown that a 0.09-inch diameter superconducting coaxial line has the same or lower attenuation for kMc waves as a $\frac{1}{2}$-inch diameter room temperature line of more complex construction.

A 400-CPS Tuning Fork Filter (O'Connor, p. 1857)—This paper introduces a new circuit component which makes use of a device that is over 200 years old. The tuning fork has long been noted as a resonance device of remarkable accuracy and stability. These properties have now been employed to produce an electromechanical filter with a narrow bandwidth and stability surpassing that which can be achieved with conventional or other electromechanical filter elements in the audio frequency range. Its application as a comb filter for detecting signals in noise, for spectrum analyzers, or for Doppler shift detectors makes it of particular interest to radar and communications engineers.

The Cyclotron Resonance Backward-Wave Oscillator (Chow and Pantell, p. 1865)—A new backward-wave oscillator has been developed which operates by virtue of electrons rotating in synchronism with the oscillations of a transverse RF electric field, thereby permitting an interchange of energy between the electrons and the field. The tube is notable because it employs so-called "fast-wave" interaction, eliminating the need for complex slow-wave structures. This structural simplification is important because, among other things, it will make it possible to scale down the tube dimensions so that it can be used in pulse operation at frequencies up to at least the one millimeter range.

Amplitude Scintillation of Radio Star at Ultra-High Frequency (Ko, p. 1871)—The twinkling and dancing of optical stars at visual wavelengths have their counterpart in the scintillation of radio stars at radio wavelengths. While it was initially thought that these fluctuations were caused by the source itself, it was later found from studies in the VHF range that they were caused by the earth's atmosphere. These studies have now been extended to the UHF range, providing significant new data regarding propagation characteristics of the atmosphere which will be of interest to many radio astronomers, space communication engineers and antenna designers.

Voltage Breakdown of Antennas at High Altitude (Scharfman and Morita, p. 1881)—The design and evaluation of missiles as well as measurements of the surroundings through which a missile travels depend on the gathering of telemetry data at ground stations. The new information presented here on how to attack the problem of high-altitude breakdown of antennas will therefore be of importance to missile systems engineers as well as to antenna designers.

A Small-RF-Signal Theory for an Electrostatically Focused Traveling-Wave Tube (Siekanowicz, p. 1888)—A complete analysis and supporting experimental data are presented for a new type of traveling-wave tube which could be of great practical interest. The tube is noteworthy in that it offers the weight-saving advantages which come with using electrostatic rather than magnetic focusing, yet it avoids the power-handling limitations associated with electrostatic systems.

Scanning the Transactions appears on page 1928.

Interrelation and Combination of Various Types of Modulation*

W. D. MEEWEZEN†

Summary—**An examination of the frequency distribution of the power in typical broadcast signals shows that low-deviation phase modulation (PM) should have advantages over amplitude (AM) and frequency modulation (FM).**

A method of stereo broadcasting is also proposed in which the sum of the two channels is transmitted as AM and the difference as PM.

Introduction

THE TYPES of modulation being considered in this paper are amplitude modulation (AM), suppressed carrier double sideband (SCDSB), single sideband (SSB), quadrature modulation (QM), phase modulation (PM) and frequency modulation (FM). While of the above types of modulation QM may not be immediately recognized as a realistic proposition, it is considered to form a logical link between the other types of modulation and, therefore, must be given its appropriate place.

It will be shown how independent sideband modulation may be treated as a special form of combined modulation, and how for certain purposes other more suitable forms of combined modulation may be employed for the provision of a twin communication channel. As an example, the problems involved in providing a twin channel for stereo broadcasting are investigated.

Interrelation of Amplitude, Quadrature and Phase Modulation

Amplitude modulation of a carrier produces two sidebands which are so phased that the resultant phase of the over-all signal is unaffected at all times. The pair of sidebands is then referred to as "symmetrical."

Quadrature modulation is obtained by a 90° phase shift of the carrier as compared with amplitude modulation. The pair of sidebands is then referred to as "antisymmetrical."

In order to study the difference between the two types of modulation, let us consider modulation by a single audio frequency.

In Fig. 1(a), OC represents the carrier vector which is amplitude modulated by the two sideband vectors CM_u and CM_l rotating symmetrically around C, thus varying the over-all signal vector sinusoidally in time between OM_a and OM_a' without affecting its phase.

A 90° shift in the relative phase of the pair of sidebands and the carrier is shown in Fig. 1(b), where the resulting signal vector varies from OM_q to OM_q'. At low modulation levels, when CM_q is small compared with OC, the amplitude of the over-all signal is virtually unaffected. Its phase, however, varies sinusoidally in time. Fig. 1(b) then approaches Fig. 1(c) where the signal amplitude remains constant at all modulation levels.

The instantaneous amplitude of OM_p is kept constant by the addition of pairs of higher-order sidebands to what is shown in Fig. 1(b). These higher-order sidebands are so phased that OM_p maintains its constant amplitude at all times.

While the above shows qualitatively how AM, QM and PM are interrelated, the relationship between the relative size of the vectors may be obtained from the mathematical expressions for the three forms of modulation.

Consider a carrier vector $OC = e_c \sin \omega t$. Amplitude modulation by a signal $S \sin \mu t$ to a modulation depth m_a then produces an AM signal:

$$S_a = e_c(1 + m_a \sin \mu t) \sin \omega t$$

or

$$\frac{S_a}{e_c} = \sin \omega t + \tfrac{1}{2} m_a \cos (\omega - \mu)t - \tfrac{1}{2} m_a \cos (\omega + \mu)t. \quad (1)$$

Quadrature modulation of the same carrier vector shifts the two sideband vectors 90° relative to the carrier. Introducing the modulation depth m_q as the ratio between CM_q and OC in Fig. 1(b):

$$\frac{S_q}{e_c} = \sin \omega t - \tfrac{1}{2} m_a \sin (\omega - \mu)t + \tfrac{1}{2} m_a \sin (\omega + \mu)t. \quad (2)$$

Phase modulation of the same carrier vector produces

$$S_p = e_c \sin (\omega t + m_p \sin \mu t), \quad (3)$$

where m_p is the modulation index which gives the maximum phase deviation in radians. This may be developed into

$$\begin{aligned}\frac{S_p}{e_c} = {} & J_0(m_p) \sin \omega t \\ & + J_1(m_p) \sin (\omega + \mu)t - J_1(m_p) \sin (\omega - \mu)t \\ & + J_2(m_p) [\sin (\omega + 2\mu)t + \sin (\omega - 2\mu)t] \\ & + J_3(m_p) [\sin (\omega + 3\mu)t - \sin (\omega - 3\mu)t] \text{ etc.} \quad (4)\end{aligned}$$

* Received by the IRE, February 15, 1960; revised manuscript received, July 11, 1960. Reprinted from *Proc. IRE (Aust.)*, vol. 2, pp. 582–590; October, 1959.

† Telecommunication Co. of Australia Pty. Ltd., Alberton, So. Australia.

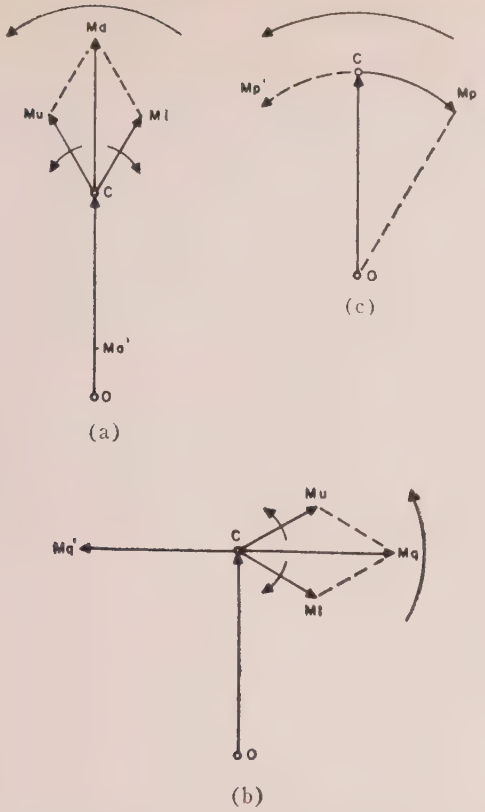


Fig. 1—Vector diagrams for (a) amplitude, (b) quadrature, and (c) phase modulation.

Here, $J_n(m_p)$ represents the Bessel function of the first kind and the nth order.

For low values of modulation index ($m_p<1$) the following approximations may be used:

$$\begin{aligned} J_0(m_p) &= \sqrt{(1 - \tfrac{1}{2}m_p^2)} \\ J_1(m_p) &= 0.5m_pJ_0(m_p) \\ J_2(m_p) &= 0.125m_p^2 \\ J_3(m_p) &= 0.02m_p^3 \\ J_4(m_p) &= 0.0025m_p^4. \end{aligned} \tag{5}$$

Fig. 2 gives an impression of what happens at higher modulation levels.

The following conclusions may be drawn from the previous paragraphs.

If the modulation depth for AM and QM is identified with the modulation index for PM, one arrives, for low modulation levels ($m \ll 1$), at identical frequency spectra. All information is then represented in the first-order sidebands which, for all three types of modulation, have the same amplitude. A modulation depth of m is then equivalent (except for phasing) to a modulation index of m.

However, at high modulation levels, ($m>1$), differences show up. Amplitude and quadrature modulation maintain their relationship in theory. In practice, AM transmitters do not permit operation at a modulation

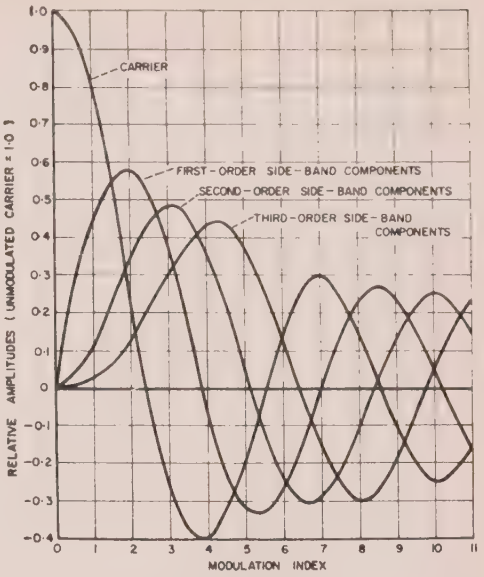


Fig. 2—Level of sideband components as a function of the modulation index for phase modulation.

depth in excess of 100 per cent which involves a phase reversal of the signal. No such problem exists with QM, while PM develops higher-order sidebands. Looking at it from a different angle, AM and QM are types of modulation whereby the sideband power is additional to the carrier power already there without modulation. For PM, however, the signal amplitude is constant and sideband power can only be obtained at the expense of carrier power. While at low modulation levels the impact on $J_0(m_p)$ is negligible, the carrier may disappear altogether at higher modulation levels (see Fig. 2).

If one compares a QM communication system and a PM system at low (identical) modulation level, the two are equivalent. By increasing the modulation level in both systems, one obtains the same (proportional) increase in receiver audio output.

The means by which this is achieved, however, is altogether different. In the QM system, the sideband power will rise with the square of the modulation level, the bandwidth of the transmission remaining constant. In the PM system, on the other hand, the total power in the transmission remains constant, and higher audio signal level is derived from increased bandwidth of the transmission.

This shows how bandwidth and transmitter power can be interchanged for obtaining a given SNR in a communication system.

Obviously, when bandwidth does not present a problem, it is advantageous to utilize a wide-band system in order to preserve transmitter power. When, however, bandwidth is at a premium, as is the case in the HF band, one must obtain from increased sideband power what cannot be achieved by increasing bandwidth. It then pays to employ a type of modulation which does not stop abruptly at a modulation level of unity, since the power contained in the carrier does not contribute

to the information transmitted. Restricted-carrier or suppressed-carrier double-sideband and single-sideband systems fall in this category and are in use. Quadrature modulation, however, deserves serious consideration.

At low modulation levels, it may be demodulated by following straightforward PM techniques. At higher modulation levels, one must make allowance for the variation in signal amplitude. A first-order approach is the omission of the AM limiter ahead of the PM detector. These limiters are normally included in order to remove any (unwanted) variation in the amplitude of the incoming signal for which the phase detector would otherwise be sensitive. If, however, the modulation is of the high-level QM type, one wants the detector output to be proportional with the product of signal level and phase deviation, and this is achieved by omitting the amplitude limiters. On the other hand, one must realize that the omission of amplitude limiting makes the receiver sensitive for amplitude *and* phase noise modulation which will reduce the efficiency of the detection system.

Frequency Spectrum of Speech and Music and its Effect on the Relative Merits of AM, QM and PM

Any assessment of the relative advantages of alternative types of modulation must take into account the nature of the intelligence to be communicated. An analysis of what is required for a single speech or music channel is most interesting from this point of view. In line-communication engineering, it has become common practice to employ standard "weighting curves" to incorporate into the calculations the statistical information available on the frequency spectrum of speech and music. A similar procedure may be followed for radio circuits. An assessment of this nature must, out of necessity, be based on statistical information. If we are to arrive at truly reliable quantitative results, a large number of "samples" must be available and the interpretation of such samples carefully watched.

In the following paragraphs, use has been made of the analysis as published by Overley.[1] This analysis provides us with the power distribution within octaves of the loud passages of various types of music and speech. It may be argued that this subdivision does not give sufficient detail because more than one tone will often appear within the one octave. Furthermore, Overley's measurements are not truly instantaneous measurements but do incorporate some short time averaging. Both of these imperfections, however, will only have a minor quantitative (favorable) effect on the outcome of the following argument.

Assuming that a loud passage causes "100 per cent" modulation level when all individual components are so phased that they add up to a maximum peak level, one may derive from Overley's information the maximum anticipated modulation levels within octaves for various types of music and for speech (Table I).

TABLE I

Percentage Contribution to 100% Modulation Within Different Frequency Bands for Various Types of Music

Frequency band (cps)	20 37	37 75	75 150	150 300	300 600	600 1200	1200 2400	2400 4800	4800 9600	9600 20000
Pipe organ	5½	20	20	14¼	12¾	12¾	9	4½	1	—
Symphony orchestra, heavy strings, woodwinds, basses	—	7¾	10¾	12	21½	15¼	24	3¾	4¼	—
Speech (male)	—	5	12½	17½	22	17½	12½	7¾	4½	1
Small concert orchestra, heavy brasses	—	9½	12	7½	15	17	21¼	12	4¼	1¼
Soprano solo with orchestra	—	1¼	3¼	7¼	26	36½	13	10¼	2½	—
Full symphony orchestra	—	3	12	4¾	24	21½	17¼	10¾	5½	1¼
Dance band, instrumental	—	4¼	9½	11	11	19¼	21¾	15½	7	1
Small string group, background, semi-classical	—	10¾	7¾	12¼	19¼	17¼	15¼	12¼	5	½
Baritone with band	—	15	16¾	15	16¾	18¾	9½	5¼	3	—
Large mixed chorus	—	2	5¼	8½	26¾	21¼	21¼	13½	1¾	—
Piano solo	—	—	6¾	17	30½	27	12	5¼	¾	—

In the above, it has been assumed that within octaves single tones are responsible for the contribution within that octave. Any alternative would make the following reasoning only more valid. It should be understood that "100 per cent" modulation has not yet been related to modulation depth or index. It is the object of this exercise to determine what this relationship should be.

Our first observation is that the highest contribution in Table I is 36½ per cent. Secondly, this highest figure only occurs below 1200 cps which means that second- and third-order sidebands would still, from an interference point of view, fall within the bandwidth of the transmission. It is only at audio frequencies above 2400 cps that higher-order sidebands might interfere with the adjacent channel assuming 10-kc channel spacing. In this region, however, our Table I puts the highest contribution at 15½ per cent. It is, therefore, unrealistic to put the maximum permissible value of the modulation index for narrow-band (first-order sideband) PM at

[1] J. P. Overley, "Energy distribution in music," IRE Trans. on Audio, vol. AU-4, pp. 120–123; September–October, 1956.

unity. Identifying a modulation index of 3 with 100 per cent modulation would restrict the modulation levels of individual components to approximately unity. The second-order sideband level within the octave 1200–2400 cps away from the carrier resulting from this modulation level is then (see Fig. 2) approximately 10 per cent which is well below the level of $3 \times 13 = 39$ per cent already anticipated from other first-order sidebands. Similarly, third- and higher-order sidebands are negligible compared with levels already anticipated as first-order sidebands of other components.

One may apply the same reasoning to the anticipated modulation level in other octaves. Again, it will be found that the higher-order sidebands fall well below first-order sideband levels. Therefore, a modulation index of 3 (or even higher) may be used in a PM system without, in practice, increasing the spectrum of the transmission.

The difference between AM and PM is that while practical AM systems cannot accept peaks in excess of a modulation depth of unity, the PM spectrum is not determined by peak levels, but by the individual components of the intelligence presented. It is true that sum and difference sidebands will occur, but these can be proved to be equally harmless.

Calculating from Table I the ratio between the anticipated peak power level and the rms power level during the loud passages, one finds that 8 db, accepted by the Postmaster-General's Department as the margin between rms and peak power level in line telephony, is a very realistic figure.

While for AM it is the peak power level which determines the point of overmodulation (with accompanying increase in bandwidth as well as distortion), it is the rms power which, for low deviation PM and AM, determines the "filling factor" of the transmission band.

For PM the cross-over from narrow-band transmission to wide-band transmission is not related to the peak power level. It is determined by the level of the individual components of the intelligence transmitted.

The above observations lead to the conclusion that, bearing in mind the statistical information on the sound power distribution in music and speech, one may permit the peak value of phase deviation to reach approximately 3 radians without increasing, in practice, the bandwidth of the transmission. Thus, assuming equally efficient detection systems, PM would be in the advantage by some 10 db as compared with AM.

The above analysis cannot be considered to be sufficiently thorough to determine what the exact value of the PM advantage over AM is. It appears evident, however, that the advantage should be at least 10 db. Sufficient evidence has been given to warrant a more elaborate investigation into the true value of the advantage, but information of this nature has not as yet become available.

Quadrature modulation would seem to offer an advantage over phase modulation in that the modulation depth can be increased beyond 3 without increasing the bandwidth of the transmission.

It seems somewhat doubtful, however, that an increase beyond 3 is of much use since at this modulation

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The level of the components contributing to the spectrum of a phase-modulated signal with n modulating frequencies may be determined as follows:

Carrier:	frequency ω	amplitude $J_0(m_1) \cdot J_0(m_2) \cdots J_0(m_n)$
First-order sideband of μ_1	frequency $\omega \pm \mu_1$	amplitude $J_1(m_1) \cdot J_0(m_2) \cdots J_0(m_n)$
First-order sidebands of μ_2, etc.	frequency $\omega \pm \mu_2$	amplitude $J_0(m_1) \cdot J_1(m_2) \cdots J_0(m_n)$
Second-order sidebands of μ_1	frequency $\omega \pm 2\mu_1$	amplitude $J_2(m_1) \cdot J_0(m_2) \cdots J_0(m_n)$
Second-order sidebands of μ_2, etc.	frequency $\omega \pm 2\mu_2$	amplitude $J_0(m_1) \cdot J_2(m_2) \cdots J_0(m_n)$
Combination frequencies μ_1 and μ_2, etc.	frequency $\omega \pm (\mu_1 \pm \mu_2)$	amplitude $J_1(m_1) \cdot J_1(m_2) \cdots J_0(m_3) \cdots J_0(m_n)$

Here, $\mu_1, \mu_2 \cdots \mu_n$ represent the individual modulating frequencies and $m_1, m_2 \cdots m_n$ the relevant modulation indexes.

It should also be emphasized that the sound power vs frequency distribution shown in Table I indicates how the anticipated maximum sideband level aids adjacent channel isolation. In Australia, the fitting of low-pass filters in the audio input circuits to broadcast transmitters is not enforced by law. Table I shows how at 10 kc or more away from the carrier, the anticipated maximum sideband level is of the order of 45 db below the carrier level in any case. Again, the results obtained in the field may be considered proof of the correctness in essence of Overley's information.

The above may be condensed into the following observations:

Speech and music both have a high ratio of peak to rms power level during the loud passages. Therefore, one must consider which power level is of significance. level already a considerable part of the transmitted PM power is contained in (first-order) sidebands. For the case of male speech and a peak value of 3 radians for the phase deviation, for example, the information in Table I corresponds with a total of 25 per cent sideband power and 75 per cent carrier power or a sideband power level only 6 db below the total radiated power level. This is approximately ten times the sideband power content available with AM. Similarly, the figures quoted for soprano solo in Table I would represent approximately 35 per cent sideband power and 65 per cent carrier power if PM with a modulation index of 3 were used. In this case, the sideband power is only $4\frac{1}{2}$ db below the level of the total radiated power.

High-level QM approaches suppressed carrier double sideband transmission. For this type of transmission,

the final stage of the transmitter must be capable of handling the peak modulation level which is approximately 8 db above the rms level. Therefore, in order to produce the same sideband power, the SCDSB or QM transmitter must be capable of handling more peak power than the PM transmitter. The power supply requirement of the latter will, however, be higher since its power output is constant.

FM vs PM

While for a pure PM system the carrier phase is deviated linearly with the modulating signal, a pure FM signal has its frequency deviated linearly with the modulating signal. The two are related through the frequency of the modulating signal as follows:

Frequency deviation = modulating frequency × modulation index.

Disregarding the frequency distribution of sound power, one would imagine that PM and FM occupy the same bandwidth when the modulation index for PM is chosen to give the same frequency deviation as the FM system at the highest audio frequency. Since the frequency deviation of FM is not dependent on the frequency of the modulating signal, its modulation index must vary inversely with the modulating frequency.

Since low-level PM has a sideband spectrum equivalent to that of AM, the random noise output of a PM receiver must be similarly "flat." This, in turn, means that FM must produce a triangular noise pattern as shown in Fig. 3.

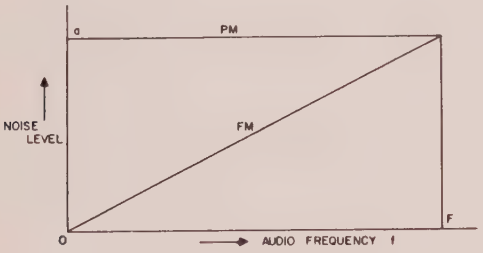

Fig. 3—Noise spectra for PM and FM.

If we assume a noise power of a^2F for the PM system, the corresponding noise power for FM will be:

$$\int_0^F \left(\frac{f}{F}a\right)^2 df = a^2F/3.$$

This would mean that the FM system has a noise level $1/\sqrt{3}$ of that of a PM system of the same bandwidth and power. Having an improvement in SNR of m for PM as compared with AM, one could quote $m\sqrt{3}$ as the improvement factor for FM. This expression is found in various textbooks.[2] For speech or music, however, it is unrealistic.

While FM has a modulation index varying inversely with the modulating frequency, PM has a constant modulation index. Thus, the frequency deviation of FM will in peaks add up to 100 per cent in accordance with Table I, but for PM the frequency deviation varies proportionally with the modulating frequency.

Assume a system employing 15 kc deviation used for the transmission of speech. The spectrum shown in Table I requires a peak phase deviation of 11 radians in order to produce a peak deviation of 15 kc. Thus a modulating frequency of $15/11 = 1350$ cps at "100 per cent" level would cause 15-kc deviation. For an FM system employing the same maximum deviation, any modulating frequency at "100 per cent" level causes 15-kc deviation. Consequently, the PM and FM systems are equivalent at 1350 cps and the curve of Fig. 3 should be replaced by the more realistic one of Fig. 4, applicable to this case.

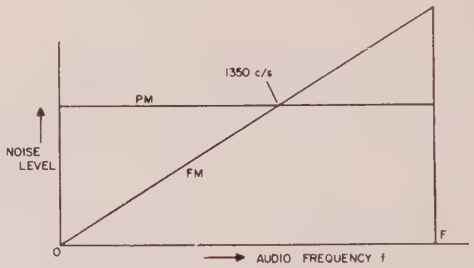

Fig. 4—"Realistic" noise spectra for PM and FM.

The above shows how, even with the receiver bandwidth cut at approximately 3 kc, PM has an SNR more than 2 db better than FM. In order to obtain the best results we use a mixture of FM at low frequencies and PM at high modulating frequencies in "FM" broadcast transmissions. The crossover point is then taken at 1500 cps.

Single-Sideband Modulation[3]

Single-sideband modulation has the obvious advantage of halving the bandwidth required for the transmission of given intelligence as compared with AM and QM. It is, therefore, of advantage whenever bandwidth is either hard to obtain or expensive. An example of the latter is found in long-haul telephone circuits where the expense of cable requires a maximum packing density of channels for economic reasons.

A second advantage of SSB transmission is that phasing of carrier and sideband for demodulation is not critical. In fact, carrier frequency errors of a few cycles are often permissible. If the carrier is not transmitted or is transmitted at a very low level, SSB offers the possibility of networking of a group of stations.

Its significant disadvantage is that it is relatively difficult to generate. At low levels and audio frequencies,

[2] S. Goldman, "Frequency Analysis Modulation and Noise," McGraw-Hill Book Co., Inc. New York, N. Y.; 1948.

[3] "Single-Sideband Issue," Proc. IRE, vol. 44, pp. 1667–1873; December, 1956.

it is possible to build very efficient SSB modulators by placing a modulator switch in between suitable filters. Power efficiencies of 90 per cent have been realized that way. At higher frequencies and high power levels, however, suitable techniques are not as yet available. The methods employed at these frequencies may be subdivided as follows:

1) the filter method whereby two first-order sidebands are generated and thereafter filters are employed to pass one and to absorb the other one,

2) the outphasing method.

The latter method is based on the following philosophy. A component $e_s \sin (\omega+\mu)t$ which forms a part of a single sideband signal may be added to any carrier, not necessarily related to the SSB signal, $2e_c \sin \omega t$.

The sum of the two signals

$$e_s \sin (\omega + \mu)t + 2e_c \sin \omega t$$

may then be rewritten as the sum of

$$e_c \sin \omega t + \tfrac{1}{2}e_s \sin (\omega + \mu)t + \tfrac{1}{2}e_s \sin (\omega - \mu)t \qquad (6)$$

and

$$e_c \sin \omega t + \tfrac{1}{2}e_s \sin (\omega + \mu)t - \tfrac{1}{2}e_s \sin (\omega - \mu)t. \qquad (7)$$

If we introduce the modulation depth $m = e_s/e_c$, expression (6) represents an amplitude modulated signal

$$e_c(1 + m \cos \mu t) \sin \omega t \qquad (6a)$$

while (7) represents a quadrature modulated signal

$$e_c\{\sin \omega t + m(\sin \mu t) \cdot (\cos \omega t)\}. \qquad (7a)$$

The outphasing method of single-sideband generation employs two sets of modulators wherein carrier as well as input are applied at a relative phase angle of $\pi/2$ and the outputs added in order to fulfil the conditions of (6) and (7). The accuracy required to cancel effectively the unwanted sideband is rather difficult to maintain.

An amplitude error between the output of the two sets of modulators of A per cent immediately shows up as a relative unwanted sideband level of $\frac{1}{2}A$ per cent. Similarly a relative phase error of ϕ radians shows up as a relative unwanted sideband level of $\frac{1}{2}\phi$.

Several means have been sought to render the unwanted sideband harmless. One of these is the so-called "third method" whereby the carrier frequency ω is chosen in the middle of the wanted sideband. This causes the residual unwanted sideband output to fall across the wanted output, but upside down. Basically it is only a special case of the outphasing method.

Combination of Various Types of Modulation

Expressions (6) and (7) show how a single-sideband signal may be considered as a combination of AM and QM with identical modulation depth and a very special phase relationship. While perfect SSB obviously has the advantage of reducing the bandwidth employed for the transmission of information, use is often made of two SSB systems, one at either side of a carrier. The whole is referred to as independent sideband modulation.

The question now arises whether one could employ AM and QM in a simpler way in order to produce two independent channels. By simplifying the procedure required for the outphasing method, one can expect to obtain more accurate results and thereby better channel isolation than what is feasible with the outphasing method.

The most awkward point of (6a) and (7a) is that the whole of the audio band must be shifted by 90° which is difficult to maintain accurately. This could be altogether avoided if one were to modulate one channel AM and the second channel QM.

Further simplification could be obtained if QM were replaced by low deviation PM. This combination can provide quite reasonable channel separation for various purposes.

Stereo Broadcasting

An interesting example is stereo broadcasting in the AM band. One is faced with the problem of providing two channels with a crosstalk requirement of approximately 20 db. The receiver complexity must be minimized for economic reasons, while the whole transmission system must be compatible with standard AM broadcasting in order to maintain the quality standards for nonaural listeners. No increase in bandwidth is permissible in view of the crowding in the broadcast band. Modifications to existing transmitters must, furthermore, be kept to a minimum.

The figure of 20-db required channel isolation may at first seem somewhat high if one only wishes to produce a stereo effect for orchestral music. Restricting stereo broadcasting to this application seems, however, unimaginative. While a channel isolation of less than 20 db brings the quality of reproduced music closer to that of the live performance, it does not take advantage of those possibilities which can be made available to individual listeners through broadcasting which, out of necessity, must be denied to the public attending the live artist performance.

A solo played in front of the orchestra provides the public with the best arrangement for a live artist presentation. Stereo broadcasting, however, opens the possibility of presenting the radio listener with the solo separate from the background of orchestral music.

The listener can then make his own decision as to how, in his environment, he will blend the solo and the orchestral accompaniment.

In view of the above it would seem desirable to aim at a channel isolation of 20 db, at least until more is known of the ways in which stereo reproduction is put to best advantage.

In deciding on a suitable mode of transmission to fulfil all of the above requirements, one must bear in mind that the operator of a domestic (variable-frequency) receiver is unskilled. Therefore, tuning errors can be ex-

pected. The equipment must, moreover be of the least complexity and of reasonable size. Its frequency response should be flat to 100 cps or lower as compared with 300 cps for speech communication systems. This rules out the filter method of ISB. The outphasing method has only disadvantages to offer as compared with AM+QM or AM+PM.

The QM content of ISB modulation is, furthermore, too high for monaural receivers employing envelope detection, the audio output signal of which would be seriously distorted unless the maximum modulation depth were restricted to approximately 50 per cent.

The requirement of providing monaural listeners with an undistorted "sum" signal may be met by transmitting the sum information of the two stereo channels as AM and the difference as QM or PM. Since, however, QM does have some second-order AM sidebands associated with it [see Fig. 1(b)], PM would be the better proposition. This combination of sum information being transmitted as AM and difference information as PM may be referred to as APM. A similar combination of AM and QM may be called AQM.

The main object of APM is to avoid distortion for monaural listeners by the difference signal transmission, which would occur if AQM were used instead. In order to avoid this "quadrature" distortion altogether for wideband receivers, one may first phase modulate a carrier with the difference signal of the two stereo channels. Following this, the signal is passed through amplitude limiters. Finally, the constant amplitude signal is subjected to amplitude modulation by the sum signal of the two stereo channels. Then the resulting envelope really represents the sum signal only.

While this method avoids quadrature distortion altogether for a wide-band AM receiver, one has to expect the vast majority of domestic receivers to have a very much restricted IF pass band. Some quadrature distortion will then be reintroduced through elimination of higher-frequency sidebands. This effect, however, causes quadrature distortion only for a very restricted range of audio frequencies. Second-harmonic distortion will occur for those audio frequencies which are still passed by the IF channel, while their second harmonics fall beyond the pass band. For a perfect IF channel, this amounts to one octave. Furthermore, combination tones may be heard from two frequencies just outside the IF pass band for which the combination difference tone falls inside the IF pass band.

In order to assess the seriousness of this effect, one should consider a narrow IF channel of, *e.g.*, 2×2.4 kc. Moreover, consider a type of music with a maximum of sound power in the higher frequencies, that is, the dance band of Table I. At 100 per cent modulation, the AM depth in the octave from 1200 to 2400 cps would be $21\frac{3}{4}$ per cent. For this octave the second-order sidebands fall outside the IF pass band. If we now assume that the PM index is kept 3 db below the AM depth, the first-order PM sidebands represent approximately 15 per cent QM. This lacks second-order sidebands equivalent to approximately $\frac{1}{2}$ per cent AM as compared with PM. Consequently, the distortion during modulation peaks amounts to $\frac{1}{2}:22=2.3$ per cent for this tone, but only $\frac{1}{2}$ per cent "equivalent modulation depth" during a loud passage.

Now consider a modulation level of $15\frac{1}{2}$ per cent between 2400 cps and 4800 cps as well as 7 per cent between 4800 and 9600 cps, causing a combination tone with an equivalent modulation depth of $2(0.7\times0.075)\cdot(0.7\times0.035)=0.26$ per cent which is of a similar order. This tone may be a difference frequency falling within the pass band of the IF channel. Difference tones of modulating frequencies within the IF pass band do not cause quadrature AM but serve to *avoid it*. Similarly, combination tones outside the IF pass band caused by modulating frequencies outside the IF pass band are harmless. It is only when some of the information relevant to certain modulating frequencies falls inside and some outside the IF pass band that quadrature distortion occurs.

Quite obviously, the frequency range of the distortion products has been reduced considerably as compared with AQM. As the pass band of the IF channel is increased, the picture will become even better. In the above case, which is an adverse combination of a narrow channel and music with high-power content at higher frequencies, the portion of the total sound power subject to (a small amount of) quadrature distortion has been drastically reduced to a figure well below 20 per cent of the total sound power.

The system does not require locked oscillators or high Q circuits for the demodulating system in the receiver. An envelope detector may be employed for AM detection, while after removal of the AM a Foster Seeley FM detector or ratio detector may be employed (with deemphasis) for the PM part of the signal. Neither of these detectors is critical in tuning. Naturally, one will lose stereo channel separation at higher audio frequencies when not properly tuned to the center of the channel. It can be expected, however, that tuning errors of up to 500 cps will not cause serious effects. Moreover, the detection system will provide a suitable source for connection of a center-zero tuning indicator which would provide a vast improvement on the present tuning indicators. Monaural listeners would hardly notice more adverse effects of tuning errors than is the case for ordinary AM transmissions. The only difference will be that detuning to one side will tend to increase the relative amplitude of one channel at higher frequencies, while detuning to the other side will tend to accentuate the higher frequencies of the other channel.

In order to achieve a channel isolation of approximately 20 db, it is sufficient that the gain of the AM and

the PM channel be equal within 20 per cent and that the phase curve of the IF channel is symmetrical within 10°. The fact that both channels pass through the same RF and IF chain in the receiver would seem to make it quite practical to keep the AM and PM gain within 10 per cent of each other. While some care may have to be taken to ensure a symmetrical IF curve, the phase curve of a critically coupled IF transformer is linear within a few degrees over its pass band. Therefore, phasing should not present a problem.

It should be emphasized that the reduction of the PM index to 0.7 for APM as compared with a possible 3 for PM transmission is brought about by the necessity of providing low distortion for receivers using envelope detection. This means that the SNR at the output of the PM detector will be 3 db lower than that of the AM detector.

The SNR of APM with a PM index 3 db below the AM depth may be calculated as follows: A tone originating from one of the stereo channels into the transmitter appears at the AM detector as well as at the PM detector. Upon recombination of the two outputs into the one stereo outlet, the two tones of equal level add up, causing a 6-db increase in signal level while the noise of the PM output is 3 db in excess of the noise level in the AM system. This causes a noise level increase of 2.4 db. Consequently, the SNR has improved by $6-2.4=3.6$ db for this channel. Allowing another 3 db for the noise in the other stereo outlet, one still finds an improvement of 0.6 db in the signal-to-noise figure. This will, in practice, be offset by the necessity to maintain a minimum carrier level of, *e.g.*, 10 per cent, in order to maintain a signal for the phase demodulator during modulation peaks of the AM system. Taking this into account, monaural listeners will lose 1 db and stereo listeners only 0.4 db signal-to-noise. Both figures would appear to be quite acceptable. A reduction of the amplitude modulation depth by 10 per cent will, furthermore, reduce the distortion in the output of the envelope detector for monaural as well as for stereo receivers. The type of APM favored is characterized by a maximum AM depth of 90 per cent and a maximum PM index of 0.64 corresponding with a maximum phase deviation of 36°. It may be denoted as APM 90/36.

Having satisfied the requirements of compatibility, channel isolation and signal-to-noise, let us now consider the increase in bandwidth resulting from the composite modulation.

A phase-modulated signal

$$e = E \sin (\omega t + m_p \sin \mu t) \tag{8}$$

which is amplitude-modulated by the same audio frequency may be written as

$$e = E(1 + m_a \sin \mu t) \sin (\omega t + m_p \sin \mu t) \tag{9}$$

or

$$\begin{aligned} e = E(1 + m_a \sin \mu t)\,[&J_0(m_p) \sin \omega t \\ &+ J_1(m_p)\{\sin (\omega + \mu)t - \sin (\omega - \mu)t\} \\ &+ J_2(m_p)\{\sin (\omega + 2\mu)t + \sin (\omega - 2\mu)t\} \cdots \end{aligned} \tag{10}$$

which may be developed (up to second-order terms only) into

$$\begin{aligned} \frac{e}{E} = {} & J_0(m_p) \sin \omega t + J_1(m_p) \sin (\omega + \mu)t \\ & - J_1(m_p) \sin (\omega - \mu)t \\ & + J_2(m_p) \sin (\omega + 2\mu)t + J_2(m_p) \sin (\omega - 2\mu)t \\ & + \tfrac{1}{2} m_a J_0(m_p) \cos (\omega - \mu)t - \tfrac{1}{2} m_a J_0(m_p) \cos (\omega + \mu)t \\ & + \tfrac{1}{2} m_a J_1(m_p) \cos \omega t - \tfrac{1}{2} m_a J_1(m_p) \cos (\omega + 2\mu)t \\ & - \tfrac{1}{2} m_a J_1(m_p) \cos (\omega - 2\mu)t + \tfrac{1}{2} m_a J_1(m_p) \cos \omega t. \end{aligned} \tag{11}$$

For a PM index of 0.7 of the AM depth, the following relationships hold: $J_1(m_p) = 0.35 m_a$ and therefore $\frac{1}{2} m_a J_1(m_p) = 0.175 m_a^2$.

Furthermore, at the low modulation levels concerned, $J_2(m_p) = \frac{1}{2} J_1^2(m_p)$ and consequently $J_2(m_p) = 0.061 m_a^2$. Also, $J_0(m_p)$ is approximately unity. Therefore, (11) may be rewritten as:

$$\frac{e}{E} = \sin \omega t + 0.35 m_a^2 \cos \omega t \tag{12a}$$

$$+0.35 m_a \sin (\omega + \mu)t - 0.35 m_a \sin (\omega - \mu)t \tag{12b}$$

$$- 0.5 m_a \cos (\omega + \mu)t + 0.5 m_a \cos (\omega - \mu)t \tag{12c}$$

$$+0.061 m_a^2 \sin (\omega + 2\mu)t + 0.061 m_a^2 \sin (\omega - 2\mu)t \tag{12d}$$

$$-0.175 m_a^2 \cos (\omega + 2\mu)t - 0.175 m_a^2 \cos (\omega - 2\mu)t. \tag{12e}$$

The two terms of (12a) represent the carrier which shifts slightly in phase; the two terms of (12b) represent the first order PM sidebands; the two terms of (12c) represent the normal AM sidebands; the two terms of (12d) represent the second-order PM sidebands which are shown to be of the AM type; and the two terms of (12e) represent introduced second-order sidebands caused by the double modulation process. These have a PM character.

Of the second-order sidebands, the ones under (12e) are the most serious ones. The total level of the two pairs of second-order sidebands at either side of the carrier is

$$m_a^2 \sqrt{0.061^2 + 0.175^2} = 0.185 m_a^2.$$

Taking the maximum figures occurring in each octave of Table I independent of type of music, it would appear that, relative to the carrier level, 1 per cent modulation between 9.6 and 20 kc may cause a radiation of −91 db in the octave 20–40 kc away from the carrier. Similarly, 7 per cent modulation between 4.8 and 9.6 kc may cause a radiation of −61 db between 9.6 and 20 kc away from the carrier. This is 17 db below the

first-order modulation products already there with ordinary AM transmission.

A $15\frac{1}{2}$ per cent modulation depth between 2.4 and 4.8 kc may cause a radiation of −47 db between 4.8 and 9.6 kc away from the carrier, which is 18 db below the first-order AM sidebands already there in the absence of PM. See also Fig. 5.

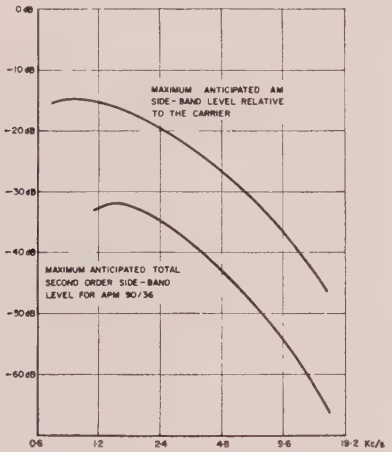

Fig. 5—Maximum anticipated sideband levels for AM and APM 90/36.

Similar results are found for any of the types of speech or music of Table I. This shows clearly that for any frequency more than 5 kc away from the carrier and during modulation peaks, the additional sidebands created are well below the first-order sidebands already there and also below the recommended figure of −40 db. Consequently, although theoretically an increase in bandwidth occurs as a result of the double modulation process, *the level of the additional sidebands is negligible by any standard.*

Possibly further improvement in the results obtained with stereo broadcasting following the APM system is possible by employing a combination of FM at low and PM at high modulating frequencies. This will introduce some problems in obtaining the correct phase characteristic in the audio circuits of the receiver. In any case, this detail does not affect the validity of the previous discussion.

The above example serves to show how for a given case the relatively simple technique of APM can provide a satisfactory mode of transmission, whereas the relatively complicated ISB technique would be unsuitable from various angles.

In general, APM or simply an independent AM and PM channel would seem the method to be employed where only restricted channel isolation of up to 30 db is required and where selective fading is not anticipated.

When, however, channel isolation of more than 30 db is required, one should not rely on outphasing, but go in for the filter method of ISB.

Conclusion

It would seem that in the assessment of the relative advantages of various types of modulation, more attention should be paid to the frequency distribution of the power contained in the information to be transmitted. The few examples contained in this study indicate that quadrature modulation and low-deviation phase modulation deserve more attention than they have been receiving in the past.

It has been generally assumed that FM and PM are unsuitable for long-distance circuits where selective fading is likely to occur [4]. This is not necessarily true for PM [5]. Early experiments appear to have been carried out with pure FM using relatively high modulation indexes. Narrow-band PM, however, employing a modulation index of up to 3, has a spectrum so closely related to AM and so very different from wide band FM that, when it is subjected to fading, one would expect a result very similar to that obtained with AM.

Such a PM communication system will have a 10-db signal-to-noise advantage over an AM system using the same carrier power.

It would, moreover, appear that independent sideband modulation is by no means the only method of providing a twin communication channel. For certain applications, such as stereo broadcasting, better results can be obtained with less complicated and less critical circuitry.

Acknowledgment

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A Categorization of the Solid-State Device Aspects of Microsystems Electronics*

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Summary—**The contribution of solid-state electronic devices to microsystems electronics is important, but there is considerable confusion in this field at the present time. By dividing the field of solid-state devices into four categories into which may be placed all device approaches to microsystems electronics (and, incidentally, solid-state components in general) we may more easily assess the merit of each new contribution by comparison with existing components. These four categories are basic, complex, integrated, and array. The construction details and operating principles, as well as the electrical characteristics, of a new device must be specified in order for it to be properly categorized. Some new examples of complex and integrated devices are presented to emphasize their differences.**

I. Introduction

MICROSYSTEMS electronics[1] is a term defining work applicable towards achieving smaller electronic equipment. It appears more descriptive of the desired end result than other proposed terms, such as microminiaturization [1], and because of this it may find widely accepted usage. Whatever term is eventually adopted, however, is the least important factor in the development of smaller electronic equipment. The technical results rank first, and will apply to any name used to describe them.

In a practical approach to microsystems electronics, it becomes necessary to consider many objectives, the more important of which are shown in Table I. None of these objectives can be appreciably compromised in the achievement of a practical operating microsystem. Items A and B are closely related, in that size and weight reductions of electronic functions will lead to increased equipment capabilities per unit weight and volume. However, there will probably be more eventual use made of microsystems electronics in providing increasingly complex electronic systems to the same (or larger) size and weight specifications than in just reducing size and weight. Increased reliability and reduced maintenance are practical necessities if more complex systems are to have a reasonable mean time to failure, and if systems in environments where repair is impossible, *e.g.*, satellites, are to function for long periods of time. Also, complex microsystems will be roughly as expensive as their macrosystem counterparts, and, if they are not repairable, long reliable service is an economic necessity.

TABLE I

Practical Objectives of Microsystems Electronics

A.	Increased equipment capabilities per unit weight and per unit volume.
B.	Size and weight reduction.
C.	Increased reliability and reduced maintenance.
D.	Improved power utilization.
E.	Reduced cost.

Improved reliability by the introduction of circuit redundancy is possible only through the reduced size of circuit components if the size of the system is to remain the same. Power must be utilized more efficiently since, with reduced size, the dissipation in conventional circuits and systems will often lead to size limitations based on thermal considerations. Low-power operation of conventional solid-state devices is not always an adequate solution to the dissipation problem because impedances often rise to unreasonably high values and operating speeds deteriorate. Also, as in macrocircuits, power must be expended to overcome the tolerances of all circuit components and to obtain the desired performance [2]. Hence, completely new electronic function approaches chosen specifically for low-power operation will often be required. Another solution is by means of components that utilize heat generated for their operation, such as thermally integrated micromodules [3] (TIMMS). This type of approach will not be discussed further here because we limit our considerations to the use of solid-state devices. Item E, reduced cost, is an obvious consideration in the practical utilization of any system.

Microsystems electronics may be applied to an electronic system at any of three levels—systems, circuits, and devices. While it is important that work in these three areas be closely integrated, the approaches may generally be set apart as shown in Table II. Although there will be differences of opinion as to how a piece of equipment performing a certain electronic function is to be broken down into systems, circuits, and devices, it must be divided in some such way for convenience in analysis, design and synthesis. Better topology and improved interconnections are technical fields being investigated by an increasingly large number of people. In general, they use some form of module upon which many circuit elements are produced or attached [4]. Modules are interconnected to produce more complex circuit functions. Novel packaging refers to the Signal

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[1] Term adopted by the AIEE Solid-State Devices Committee, subcommittee on Microsystems Electronics.

TABLE II

Approaches to Microsystems Electronics

Systems
- Novel functional organization.
- Circuit developments.

Circuits
- Novel circuitry.
- Better topology.
- Improved interconnections.
- Device developments.

Devices
- Novel packages.
- New devices.

TABLE III

Categories of Solid-State Electronic Devices

I. Basic.
II. Complex.
III. Integrated.
IV. Array.

TABLE IV

Basic Device

The simplest useful device exhibiting a basic solid-state phenomenon.

Examples
- Thermistor (− or + temperature coefficient) [5]
- Magnetoresistor [6]
- Hall effect generator [6]
- Magnetic barrier layer rectifier [6]
- Piezoelectric element [7]
- Nonlinear inductor [7]
- Electroluminescent element [8]
- Impact ionization switch [9]
- Cryosar [10]
- Superconductor [11]
- Photoconductor [12]
- Photoemitter [5]
- Infrared detector [12]
- Phosphor [5]
- Switching core [13]
- Thermoelectric element [14]
- Nonlinear resistor [5]
- High-conductance diode [15]
- High-voltage diode [15]
- Voltage reference diode [15]
- Varistor diode [5]
- Nonlinear capacitance diode [16]
- Accumulation diode [17]
- Exhaustion diode [17]
- Tunnel diode [18]
- Backward diode [18]
- Oscillistor [19]
- Pulse injection delay element [20]
- Floating junction photodetector [21]
- Avalanche electron emitter [22]
- Strain gauge [23]

Integrated Device

A device containing within its structure more than one recognizable basic or complex device, in which the interconnection medium is used only for electrical connection or isolation.

Examples
- Solid circuits (multivibrator, etc.) [47]
- Cryosar matrix [10]
- Double emitter transistor [48]
- Resistance-capacitance distributed elements [49]
- Unipolar transistor matrix [41]
- *p-n-p-n* switch-resistor combination [50]
- Tunnel diode-resistor combination [50]
- Notched filter [51]
- Diode matrices [52]

Complex Device

A device containing within its structure more than one recognizable basic or simpler complex device, utilizing the properties of the interconnecting medium to perform a function other than electrical connection or isolation.

Examples
- Bipolar transistor [5]
- Unipolar transistor [5]
- Filamentary transistor [5]
- Avalanche transistor [5]
- Unijunction transistor [24]
- Drift transistor [5]
- Analog transistor [20]
- Hook collector transistor [25]
- Remote base transistor [26]
- Punchthrough-avalanche transistor [27]
- Spacistor [28]
- Deplistor [29]
- Field effect varistor [30]
- Avalanche injection diode [31]
- Tetrode-base bias [5]
- Emitter tetrode [32]
- Unipolar tetrode [33]
- Pentode transistor (and multiple switching device) [34]
- Photodiode [5]
- Solar cell [5]
- Photoelectromagnetic (PEM or PME) cell [6]
- Photorectifier [35]
- Injection electroluminescent diode [36]
- Junction Hall device [37]
- *p-n-p-n* switch [38]
- *p-n-p-n* thyristor [39]
- *p-n-p-n* stepping element [40]
- Thyristor stepping switch [41]
- Hook collector binary full adder [42]
- Piezoelectric transformer [7]
- Compensated avalanche diode [43]
- Shorted emitter switch [44]
- Tunnel emitter transistor and switch [45]
- Cryotron [46]
- Electroluminescent-photoconductor element [8]

Array Device

A multitude of similar basic, complex, or integrated devices, without separate enclosures, each having at least one of its electrodes connected to a common conductor, or all connected in series.

Examples
- Cryotron array [46]
- Ferrite memory plate [13]
- Electroluminescent-photoconductor array [8]
- Electroluminescent display panel [8]
- Solar cell panel [53]
- Tunnel diode array [18]
- Thermoelectric generator array [54]

Corps Micromodule program, for example, in which the main contribution is a change in shape, size, or interconnection mode of the package.

The development of new processes for making materials and devices can be an important part of microsystems electronics. These include sputtering for making thin film resistors and capacitors. Certain other processes are of great scientific and commercial interest, but of questionable necessity in applications aimed specifically at microsystems electronics in the foreseeable future. These include dendrite growth and evaporation of semiconductor single crystal thin films.

The area to be specifically covered in this paper is that of solid-state electronic devices. This field is the one in which perhaps most is to be gained in microsystems electronics, and is certainly the field to which most publicity has been given. This aspect of microsystems electronics is also the one in which most confusion exists. The purpose of this paper is to categorize solid-state electronic devices along functional lines such that the contribution of a new, or existing, device-approach to microsystems electronics (and incidentally to solid-state technology in general) can be better appreciated, and its relationship to other devices more easily seen.

II. Solid-State Electronic Device Categorization

We divide the field of solid-state electronic devices into four categories, as shown in Table III. Definitions for each category are given in Table IV, along with illustrative examples. The list of examples in each case is not complete, but is made rather large to indicate the diversity of solid-state devices that now exist (even though many of them are not being manufactured). In order to assess properly a new device, it is necessary to examine it with respect to all the other devices of similar operation or structure about which information has been published. Hence, the proposed system of categorization demands that a new device be described not only in terms of its characteristics but also in terms of its construction and principles of operation. A "black box" device performing a certain electronic function cannot be judged realistically because there are generally so many ways in which various functions can be performed using the multitude of different solid-state effects, that such a black box could contain any one of a large number of devices or device combinations.

A basic device is defined as one exhibiting a solid-state phenomenon in a simple, useful way. Since a solid-state device can seldom, if ever, be made to exhibit only one physical effect, the phenomenon considered is the one primarily responsible for the major characteristic of a basic device. Space does not permit a discussion of each device, but the interested reader can obtain further information from the references cited. Some of these devices have been recently placed in production (thermoelectric element, tunnel diode), while others have been available for many years. Some devices have found limited usefulness outside the laboratory (Hall rectifier, pulse injection delay element) and others work via as yet incompletely explained mechanisms (cryosar, oscillistor).

A complex device generally has a more complicated internal structure than a basic device, and can always be mentally divided into two or more basic devices. For example, the (bipolar) transistor may be considered as composed of a low-impedance minority-carrier emitting junction (high-conductance diode) and a high-impedance minority-carrier collecting junction (high-voltage diode) connected by a base region that performs the essential function of allowing an appreciable fraction of the injected carriers to travel from the emitter junction into the collector junction. The two diodes, made separately and then hooked together with wires, would not display transistor action. The complex devices listed range from those in mass production (bipolar transistor, *p-n-p-n* switch) to those that have not found commercial importance (remote base transistor, deplistor).

Differences of opinion as to which category, basic or complex, best describes certain devices are expected, since in some cases the state of understanding is rather limited and also since many solid-state phenomena are exhibited to various degrees of prominence in each device. However, these two categories taken together cover most of the sophisticated device innovations. They share the common property that no basic or complex device can be physically subdivided into simpler basic or complex device sections, and then be connected together with wires to perform the same electronic function in the same way.

Integrated and array devices may be looked upon as being made up of separate basic or complex devices interconnected in particularly convenient ways. In an integrated device, the structure is made within a continuous piece of solid (monolithic structure), but it would work in exactly the same way, or in many cases better, if it were subdivided into its basic and complex component parts, and then reassembled with wires or other forms of connection. The first example in the list of integrated devices, the solid circuit, is a term that could be used to describe this entire category. Which term, integrated device or solid circuit, is used will depend upon the viewpoint of the user.

The array device consists of separate similar devices connected in a simple series of parallel arrangement. The difference, for example, between a diode matrix and a diode array is that in the former case all diodes would be made on the same piece of semiconductor, while in the latter case they would all be separated, the only connection being through metallic conductors. An array device may, like the integrated device, be called a circuit, depending upon the viewpoint of the user.

The advent of a new basic or complex device will almost always have a much greater impact upon microsystems electronics (as well as upon electronics in general) than a new integrated or array structure. This

is because the former constitute the building blocks for the latter. The generation of new integrated and array devices depends primarily upon suitable process control, but an innovation in basic or complex devices constitutes invention or discovery, and hence cannot be scheduled to take place within a certain period of time.

The present categorization of solid-state electronic devices is one that should be usable well into the future, since it is capable of absorbing all foreseeable types of new developments as they occur. Certain changes in classification of individual devices will occur from time to time as better understanding indicates the necessity for them. This scheme does not require use of a term like "functional block" or "functional electronic block." A functional block is generally defined as a new device the characteristics of which can be reproduced only through the use of a plurality of existing devices, or, alternatively, as a device performing an electronic function. However, by the first definition, almost every solid-state electronic device (all those in Table IV and more), at the time of its inception, might have been called a functional device since its electrical characteristics could have been reproduced only through the use of many previously known devices. Hence, in this sense, the adoption of the term functional block will do nothing more than establish a point in time, before which we shall have had devices, and after which we shall have functional blocks.

Some recently publicized devices perform what is considered an electronic function (amplifier, oscillator, multivibrator, telemeter) and so, by the second definition, would qualify as functional devices. From a practical standpoint, however, there is little agreement as to what constitutes a function. For example, a transistor is an amplifier, with bias supplies and load, but a practical amplifier is generally required to provide (in addition to gain) stability, bandwidth, linearity, impedance matching, etc. Amplifiers with a great variety of complexity are in use at the present time. As further characteristic requirements are added to those of the simple transistor, at which point does it cease to be a device and commence to be a functional block?

The term "molecular electronics" (molecular engineering, molectronics) is defined as the design of appropriate materials and the provision of appropriate electrodes to produce a desired electronic effect [55]. Molecular electronics, then, differs from research and development in solid-state materials and devices for the reason that in molecular electronics the electronic function is the reason for, not just the result of, the research and development effort. Nevertheless, we should be able to place any development emanating from a molecular electronics program into one of the device categories—basic, complex, or integrated. The publicized examples of molecular electronics [50], [51], [56], with few exceptions, if any, fit logically into the integrated device category. The practical advantages of integrated devices in microsystems electronics include extremely small size, the elimination of many, or all, wire interconnections, and perhaps some cost reduction. In some cases, these advantages outweigh the disadvantages, and the resulting devices will be generally adopted. However, for the benefit of the reader not intimately connected with solid-state devices, it is in order to list the limitations and potential disadvantages of the integrated-device approach to microsystems electronics so that a more practical comparison with the other approaches may be obtained.

Integrated structures should, inherently, have no better device reliability than circuits using the separate devices because the actual number of devices has not been reduced. The reduction in the number of wire interconnections by using metallic regions on the surface can help reliability only if the wire contacts cannot be made as reliably as the metal layers and their crossovers. Many circuits cannot be laid out in a geometric pattern suitable for inclusion in a simple integrated form. In such cases, the pattern of surface metallic regions will be much more complex than the required wire interconnection scheme, and interconnection reliability might be worse than with wires. In the integrated structure, it is generally impossible to select component devices according to their electrical characteristics, as is commonly the case when components are manufactured separately. Hence, tolerances will tend to be wider, giving poorer circuit performance or requiring greater power dissipation for equivalent performance. It is seldom possible to replace a bad device in an integrated structure, and one bad component will usually make the entire integrated structure inoperative.

The above raises questions concerning yield and cost. The integrated device is limited technologically in the range of materials which may be used in a single structure. This limits function versatility where one might have reason to use, by way of illustration, a diode, transistor, tunnel diode, capacitor, thermistor, and resistor, each made of a different material, to obtain superior electrical characteristics. In the integrated structure, practical process and geometrical limitations make the exact reproduction of most circuits impossible. This results in the appearance of parasitic resistances between the desired devices, increasing power dissipation above that in the same circuit made using separate devices, and degrading circuit performance. It is often impossible, in an integrated structure, to test for the electrical characteristics of each component, basic or complex device, let alone their stability. The over-all characteristics of the integrated device are measured, and its reliability ascertained. This is the equivalent, in conventional circuitry, of making the particular electronic circuit using untested components, measuring the over-all functional performance, and discarding those units that do not meet performance or stability specifications. In such a case, it is impossible to design the circuit for "worst case" component values; hence, there is required much testing of the completed circuit to establish reliability,

as is now conventionally done with individual basic and complex devices. This can be done economically with a particular integrated device only if it is produced on a large scale.

III. Some New Examples of Complex and Integrated Devices

Some examples of complex and integrated devices will be described in detail to illustrate the differences in definition between these two categories, and to emphasize the necessity of knowing the structure of a new device before it can be adequately judged or categorized. The examples are new in structure, performance, or in explanation of operating mechanism. New examples are chosen, rather than older ones, for added interest to those already familiar with the field, as well as to emphasize the common (and random) occurrence of solid-state electronic device innovations.

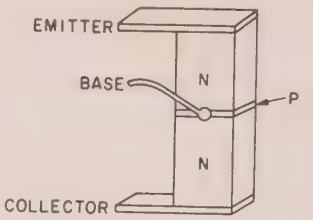
A. Complex Devices

As defined in Table IV, a complex device contains, within its structure, more than one recognizable basic or simpler complex device, in which the medium interconnecting these "internal" devices performs a function vital in the operation of the over-all device, other than electrical connection or isolation.

1) Controllable Short-Circuit Stable Negative Resistance: A bar-type transistor with a small wire contact to the base region exhibits crowding of emitter current about the base contact. The degree of crowding increases with base sheet resistance, larger degrees of crowding resulting in lower current gains caused by greater surface recombination losses of minority carriers injected from the emitter to the base. When the transistor is in saturation, both junctions are forward biased so that, at sufficiently high currents, base conductivity modulation takes place. The resulting lower base sheet resistance allows the emitter current to spread more evenly across the emitter than when the transistor is out of saturation (and the base conductivity modulation is much less). Hence, a drop in current gain takes place as the transistor comes out of saturation, producing a short-circuit stable negative resistance that is controllable by means of the base current. This is illustrated in Fig. 1. Like the transistor, this structure cannot work in the same way if physically subdivided into simpler basic or complex component parts and wired together.

2) A Tunnel Unijunction Transistor Threshold Resistance: As shown in Fig. 2, a tunnel junction, unijunction transistor can be readily fabricated. A degenerate n-type region (designated $n+$) is diffused into the bar, and a degenerate p-type region is formed by alloying, making a tunnel junction ($p_t n$). The measured resistance is between bases 1 and 2, and the control current enters at the p-type contact. For junction currents below the peak current of the $p_t n$ junction, the interbase resistance has its unmodulated value. In this region the tunnel current, flowing through the upper or lower part of the bar (depending upon polarity of the base 1-base 2 bias) modifies the potential distribution within the bar and hence contributes to the apparent interbase resistance. When the current at the tunnel junction exceeds the peak current, the tunnel emitter switches to the normal thermal region and emits minority carriers into the bar. The interbase resistance then drops abruptly, because of resistivity modulation in the base region.

3) Ring Base Transistor dc Switch: When some ring base alloy transistors are driven to high currents in the avalanche region, with a reverse-biased emitter junction, they switch to a low voltage state [57]. This is a dc state, not to be confused with the transient switching "avalanche mode" [58]. Fig. 3 shows the electrical characteristics, the circuit, and the device space-charge geometry at various stages of bias for a properly



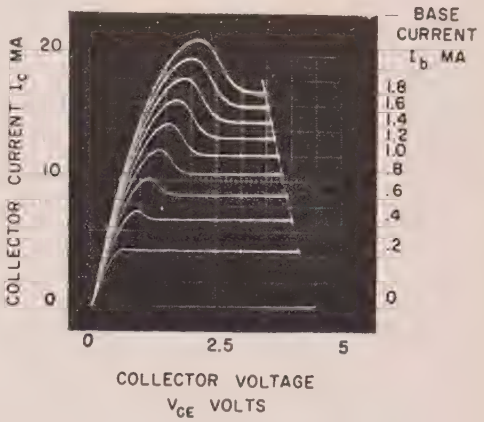


Fig. 1—Controllable short-circuit stable negative resistance.

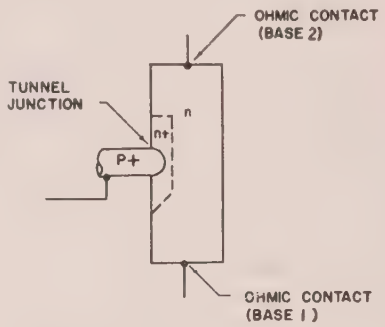


Fig. 2—Tunnel unijunction transistor threshold resistance.

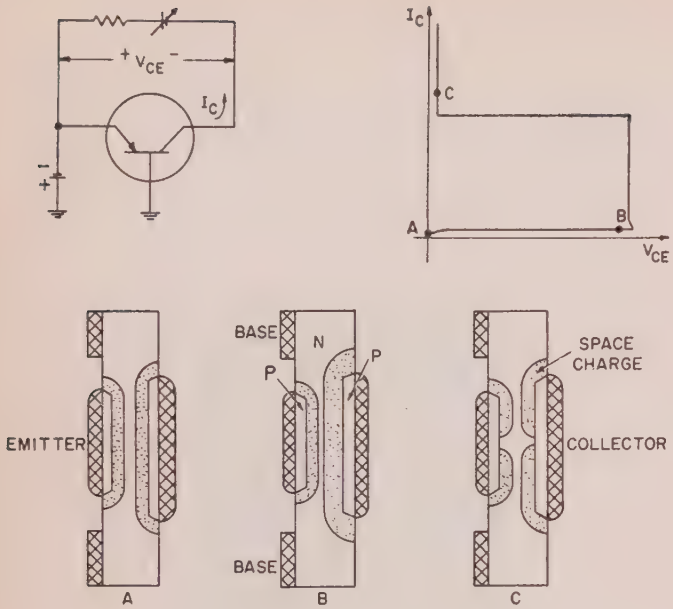


Fig. 3—Ring base transistor dc switch.

designed switch. Space-charge condition *A* is for the collector shorted ($V_{CE}=0$). As V_{CE} is increased, the collector space-charge widens; that about the emitter remains the same because of the constant V_{EB}. Avalanche occurs when the space-charge regions have almost touched, condition *B*. The electrons released into the base by the avalanche leave as drift current to the base contact. Since the lateral base resistance is high due to the space-charge pinching, electrons accumulate in the center of the base, neutralizing the positive space charge there, and thus allowing a large current to flow from emitter to collector at a low voltage, condition *C*. The constant reverse bias applied between the emitter and base appears across the space-charge "channel" region. In the low-voltage high-current state, this device utilizes bipolar and unipolar transistor action, but the characteristic cannot be reproduced by a separate unipolar and bipolar transistor hooked together.

4) Light Transistor: The operation of this device is predicted from the operation of other devices. The light transistor is a device similar to a bipolar transistor, in which energy is transported across the base by means of light (or other electromagnetic radiation). The emitter is a low-impedance injection electroluminescence junction. Light produced in the base travels in all directions, but by proper device design, a reasonable per cent will intercept the collector high-impedance *p-n* junction. The collector region is of a smaller bandgap than the base, or contains appropriate impurities so that it has a large absorption coefficient for the radiation coming across the base. Since the absorbed radiation creates carriers in the collector in direct proportion to its intensity, the collector current is proportional to the emitter current. If the collector junction impedance is much higher than that of the emitter, power gain should be possible.

B. Integrated Devices

As defined in Table IV, an integrated device contains within its structure more than one recognizable basic or complex device, in which the medium interconnecting these "internal" devices performs no function vital to the operation of the over-all device other than electrical connection or isolation.

1) P-N-P-N Switch—Junction Capacitance Integrated Device: The ordinary two-terminal *p-n-p-n* switch has, at a given temperature, a stable negative resistance region. Properly biased, and with sufficient external capacitance across it, the device will oscillate in a relaxation mode. A reverse biased *p-n* junction of sufficient area will, of course, serve as the external capacitance. A possible structure is shown in Fig. 4. If the switch is so designed that little multiplication from the junction J_c is required for breakover to occur, the leakage through J_{cap} may be assumed to remain small enough so that the capacitance of the junction is its most important parameter; relaxation oscillations should be possible. The "shorted emitter" principle has been applied to many variations of *p-n-p-n* switches, generally for the primary purpose of stabilizing the breakover voltage [44]. A shorted emitter type structure, analogous to that of Fig. 4, is shown in Fig. 5. Bistable operation of the device is possible, but, in general, the portion of the junction J_c external to the *p-n-p-n* structure is forced into a region of high multiplication before "firing" occurs. If the external area is very large, a large firing current is required, there is no stable negative resistance region, and no relaxation oscillations are observed.

If, however, the avalanche voltage of the external portion of J_c is made much higher than that of the portion within the *p-n-p-n* structure, the switch will fire at currents less than the hold current. The voltage stabilization of the shorted emitter structure will be retained (since avalanche is still required to produce the current necessary to bias on the emitter), but the external junction will remain essentially a capacitance. A stable self-contained oscillator, requiring merely a suitable bias supply (and series-charging resistance) is thus obtained. Frequency of oscillation for a given device is a function of the charging resistance, which can be a photoconductor if frequency modulation by light should be desired. Since the *p-n-p-n* section breakdown voltage is decreased and leakage current is increased by photoinjection, light modulation of frequency may be obtained by shining it on the junction structure itself, rather than on a series photoconductor. Another modification of the structure is to use punchthrough instead of localized avalanche to produce the desired negative resistance characteristic.

Devices with undescribed but apparently similar internal construction have been called ROOT systems in the literature [51].

2) Darlington (Compound) Transistor Integrated Con-

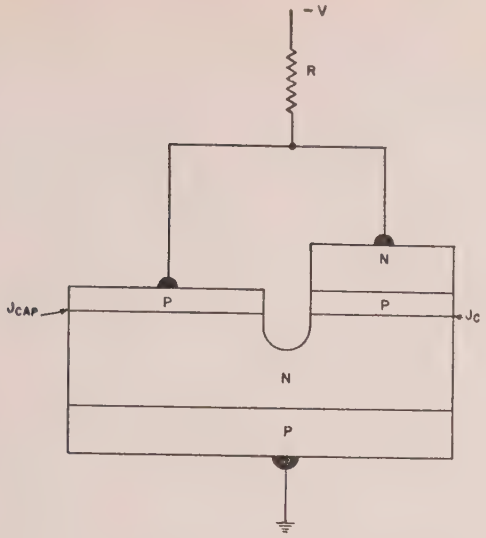


Fig. 4—*P-N-P-N* switch—junction capacitance integrated device. Separate section construction.

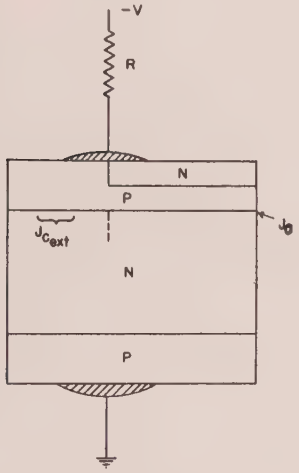


Fig. 5—*P-N-P-N* switch—junction capacitance integrated device. Shorted emitter construction.

nection [34]: The fact that a shorted emitter, remote from the short, can act in a more or less normal manner, makes possible the easy construction of a two-stage (or more) direct coupled amplifier, employing the Darlington circuit in the integrated form of Fig. 6. Devices made this way have shown current gains much higher than for single devices made from the same structures, and should approach the product of gains theoretically possible if the two transistors were made separately and wired together. For best performance, the resistance of the base region between the short and the input circuit should be high so that the effect of the short on the current gain of the first stage will be minimized; *i.e.*, parasitic conductance between bases of the transistors should be low. Also, the ratio of sizes of the two sections should be roughly proportional to the current gain so that both sections can operate at the optimum current densities. This requirement suggests a

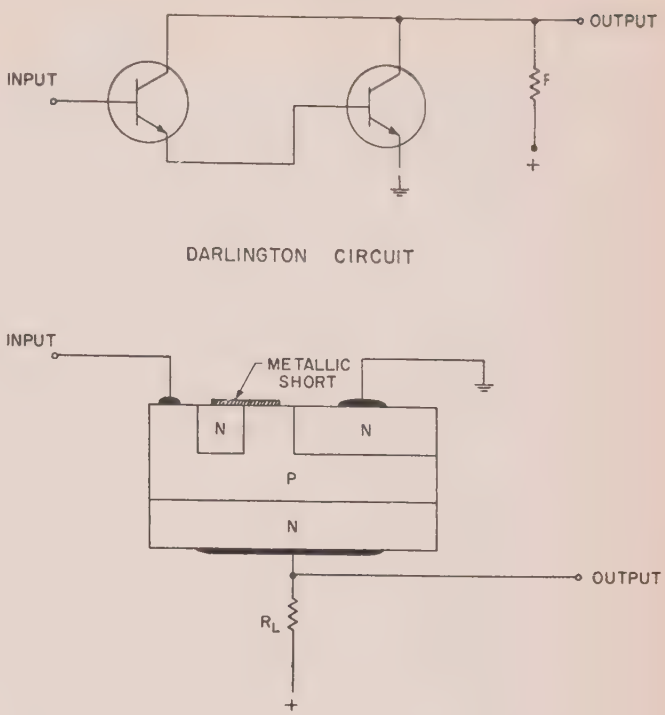


Fig. 6—Shorted emitter structure equivalent to Darlington circuit.

circular geometry with the input section in the center and a concentric output section. It is apparent that equal or better performance would result from the two separate transistors wired together.

Devices of this type, with undescribed but apparently similar internal construction, have been called amplifiers in the literature [56].

3) Solar Battery-Tunnel Diode Integrated Oscillator: An example of an integrated, light-driven, tunnel diode oscillator structure is shown in Fig. 7(a) with its equivalent circuit in Fig. 7(b). The structure consists of a single crystal slab with a light-sensitive, large-area diffused junction at the top of the crystal wafer and a degenerate diffused layer upon the lower face of the wafer. A tunnel junction is alloyed to the degenerate diffused layer and is connected inductively directly back to the upper face of the wafer, thus completing a series circuit consisting of solar battery, bulk wafer resistance between solar junction and tunnel junction, tunnel junction and associated capacitance, and the inductive connection itself.

The design of the structure must be such that the circuit resistance is less than the absolute value of the tunnel junction negative resistance. The inductive connection can be a simple wire or evaporated metallic connection from top junction to lower tunnel junction. The inductance of the loop may be cut to a very low value by alloying the connecting inductive link directly through the wafer as shown in Fig. 8.

Structures of the type shown in Figs. 7 and 8 have been fabricated in *n*-type silicon which was first heavily diffused with phosphorus or arsenic (preferably arsenic). One side of the wafer was then removed by grinding

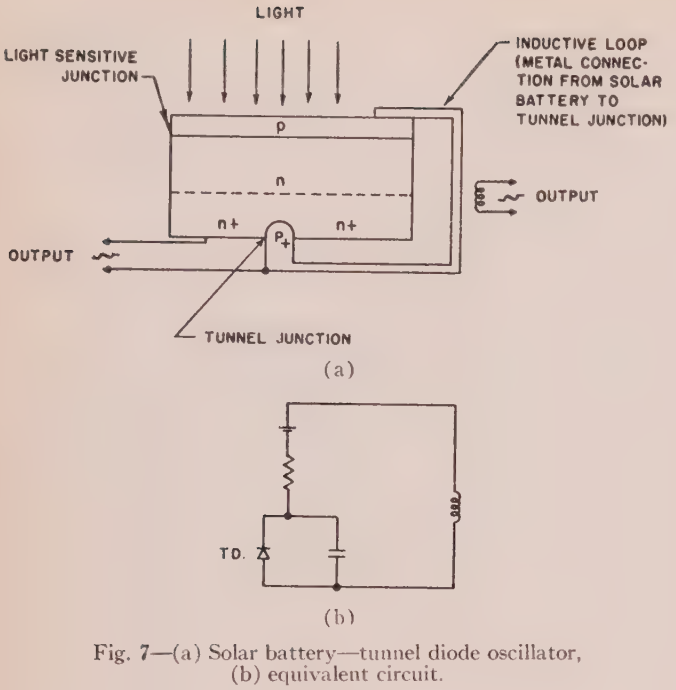


Fig. 7—(a) Solar battery—tunnel diode oscillator, (b) equivalent circuit.

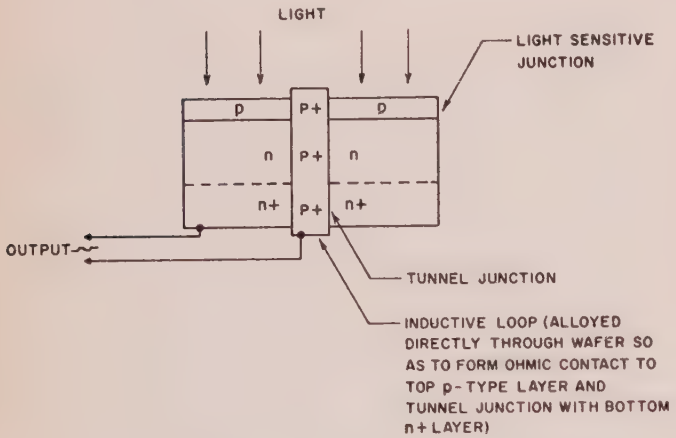


Fig. 8—Solar battery—tunnel diode oscillator with internal inductive loop connection.

and etching, and the opposite side was oxide-masked prior to boron diffusion of a light-sensitive junction into the etched side. Finally, a tunnel junction was alloyed to the heavily phosphorus or arsenic-doped side of the wafer, and the inductive connection to the diffused junction was completed. Light-produced oscillations were obtained at 10 Mc when approximately a two-foot length of wire was used as the inductive connection (Fig. 7). Oscillations have not yet been observed from the integrated device shown in Fig. 8. In either case, it is apparent the structures could be made separately with a tunnel diode and a solar cell.

4) Unijunction Transistor—Capacitor Integrated Device: The unijunction transistor, shown in Fig. 9, is another device that is amenable to making integrated, oscillating structures. If the device is made, for example, from germanium, its junction terminal floats in the middle of a negative resistance region at room temperature [59]. A capacitance placed between the junction and base 1 results in relaxation oscillation. The capacitance may be in the form of an oxide on the *n*-type region surface, or an extra-integrated junction structure. The frequency of oscillation may be tuned by light or other absorbed radiation over a large frequency range.

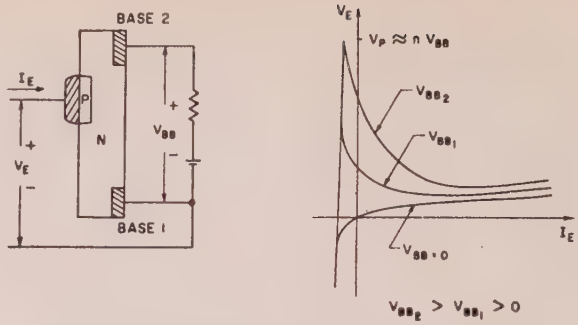


Fig. 9—Unijunction transistor and its negative resistance characteristics.

IV. Conclusions

The field of solid-state electronic devices has been divided into four categories—basic, complex, integrated and array. Each existing and new device may be placed in its proper category provided that its operating characteristics and structure are described. Each solid-state approach to microsystems electronics may likewise be placed in its logical category. Use of this system permits a simple, accurate, unemotional comparison of all new solid-state device developments, and will, hopefully, help in planning for the best future directions to be followed in microsystems electronics.

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A Two-Color Input, Two-Color Output Image Intensifier Panel*

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Summary—**The design and construction of a two-color input, two-color output image intensifier panel are described. The solid-state panel is so constructed that it can be independently excited by either a visible image or a near-infrared image on the input. The visible input image excites one of the colors in the output image while the infrared input image excites the other color of the output image. The characteristics of the panels are given in terms of input and output light levels, spectral response and gain. Photographs are given which show the output image for visible input only, infrared only, and the two combined.**

Introduction

THE ADVENT OF electroluminescence as a new solid-state light source, and the production of highly sensitive photoconductive powders have in the last few years made possible the fabrication of solid-state image intensifier panels or light amplifiers. These first panels were monochrome devices giving images in one color when excited by a source with a spectral range within the response of the input photoconductor. Two photoconductive powders with considerably different spectral responses are now available, and at the same time electroluminescent phosphors of more than one color have been developed. These two advances have now made possible the construction of a two-color input, two-color output image intensifier, although such a device has also introduced new problems in large-area panel construction.

Large-area solid-state light amplifiers have been described in several publications,[1,2] and, in particular, the details of the photoconductive-electroluminescent type have been emphasized. In general, the amplifier consists of a layer of photoconductor on a layer of electroluminescent material with an opaque layer between the two to prevent optical feedback. Alternating voltage is applied between the electrodes on the outer surface of the photoconductor and the under surface of the electroluminescent layer, thus forming a series circuit of photoconductor and electroluminescent material in which current flow through the electroluminescent layer is controlled by the photoconductive layer resistance. This resistance is in turn controlled by light falling on it in the form of an incident image. Because of the intimate spatial relation between the photoconductive and electroluminescent layers, these variations in resistance of the photoconductor appear as variations in light from the output electroluminescent layer. If the sensitivity of the photoconductor is great enough, then the output picture is brighter than the input.

Several types of monochrome light amplifiers have been described by Kazan and Nicoll[1,2] but of these, the grooved photoconductor type seems most suitable for the present application. It has the advantage of high gain, high resolution and a fabrication technique which has been fairly well worked out. The use of grooves in the photoconductive layer obviates the difficulty that light does not penetrate the photoconductive powder to any appreciable depth. It is also of great advantage when more than one photoconductor is to be used as in the case of a two-color device.

Two-Color Input, Two-Color Output Light Amplifiers

In order to make an image intensifier or a light amplifier which is sensitive to input images of two or more discrete colors and will reproduce these in two separate color output images, certain structural modifications in the design of monochrome amplifiers are required. In one form of two-color device it is necessary to divide the area into discrete small color areas on the output, and so register these with the input that excitation by one color at the input will excite only one color at the output, and excitation by another input color will excite only a second color at the output. For ease of construction in a two-color system the discrete areas are preferably in the form of strips. This makes input-output registration relatively easy and in addition fits in particularly well with the grooved structure already worked out for monochrome light amplifiers.

Although it is possible to separate images on the input and output by the use of appropriate filters, suitable filter characteristics are difficult to obtain and considerable transmission loss would have to be tolerated. For this reason, and with the availability of suitable materials, the present panel was made with two photoconductors on the input and two separate electroluminescent phosphors on the output.

Early work on materials for light amplifiers produced two sensitive photoconductive powders with rather well-separated spectral response peaks. These two ma-

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terials are cadmium sulfide and cadmium selenide.[3] The spectral response of these two materials is shown in Fig. 1 along with the relative output of suitable image sources referred to later. The cadmium sulfide powder is a brown material and has a broad visible sensitivity extending from 500 millimicrons to 800 millimicrons as shown in curve A. A typical cadmium selenide powder, which is black in appearance, has a response between 700 and 1000 millimicrons as shown in curve B. The spectral peaks of the sulfide and selenide are not sufficiently separated to respond independently to visible and infrared excitation unless the visible input is restricted in its infrared content. This is the manner in which the panel is operated.

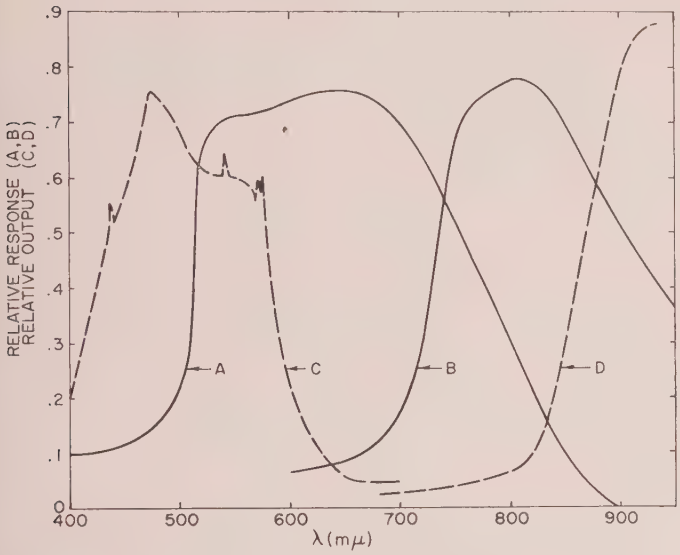

Fig. 1—A. Spectral response of cadmium sulfide powder. B. Spectral response of cadmium selenide powder. C. Spectral output of fluorescent-lamp, "visible source." D. Spectral output of filtered tungsten light, "IR source."

In addition to the use of two photoconductive materials on the input side, two different electroluminescent materials are used on the output side. There is now available a choice of electroluminescent materials with a variety of emission colors. However, for this particular application maximum color contrast is desirable, but at the same time the materials should be comparable in brightness. For this reason a blue emitting phosphor RCA F2102, and a yellow emitting phosphor RCA F2099, were chosen as most suitable.

Fig. 2 shows a diagrammatic section of the two-color light amplifier with the various dimensions of the layers. It shows the alternate ridges of plastic-embedded cadmium sulfide and cadmium selenide powder with the conducting silver lines on top. Each ridge is registered with a strip of the appropriate color of electroluminescent layer, and feedback of output light to the input is prevented by an opaque layer under each photoconductive ridge. There is also a current diffusing layer of non-photoconductive cadmium sulfide powder, RCA F2108, which has a nonlinear conducting characteristic. This property assures slight sideways spreading of the photocurrents without loss of resolution. It has the value in this particular design of causing photocurrents from a particular photoconductive ridge to spread under the ridge, thus exciting the entire width of the phosphor strip directly beneath the ridge. Spreading of currents from one ridge to the adjacent phosphor strips is prevented by nearly cutting through the current diffusing layer and thus dividing it into strips which are electrically isolated from each other.

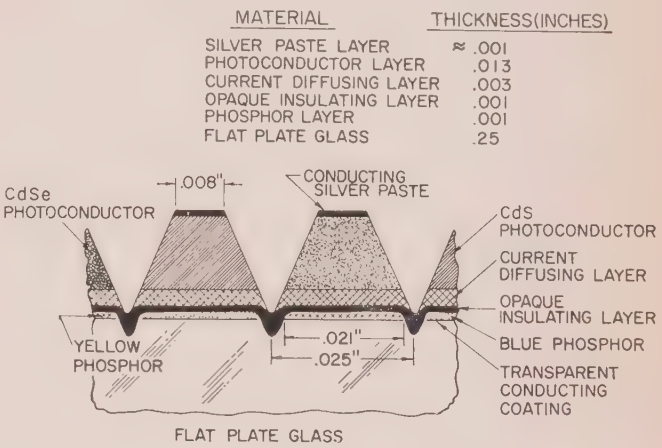

Fig. 2—Diagrammatic section of two-color input, two-color output light amplifier.

Construction

Construction of the two-color light amplifier begins with the selection of a piece of glass with faces flat and parallel. "Parallel-O-Plate," manufactured by Libby Owens Ford, is suitable. One edge of a 6 inch×6 inch sheet is ground square to serve as a reference edge for subsequent machining operations. One face of the glass, except for two $\frac{1}{4}$-inch bands at the reference edge and at the opposite edge, is covered with a transparent conducting coating. The next step is application of the one-mil thick electroluminescent layer, in this case consisting of alternate strips of plastic-embedded yellow and blue phosphor parallel to the reference edge. These strips are almost twenty-five mils wide, and are spaced on twenty-five mil centers. Fig. 3 shows how this is accomplished. First, twenty-five mil wide strips of aluminum are evaporated over the transparent conductive coating (TCC), spaced with centers fifty mils apart and parallel to the reference edge [(Fig. 3(a)]. Then a layer of blue electroluminescent phosphor in plastic binder is sprayed on and baked to cure the binder [Fig. 3(b)]. After the bake, the phosphor is lapped through using a brass tool whose teeth are twenty-five mils apart; the cuts are made so that the strips which have aluminum

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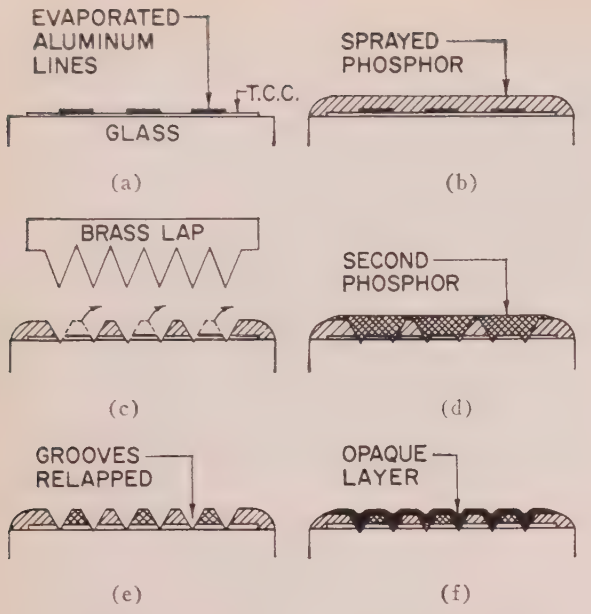

Fig. 3. Steps in construction of two-color electroluminescent output layer (vertical scale exaggerated).

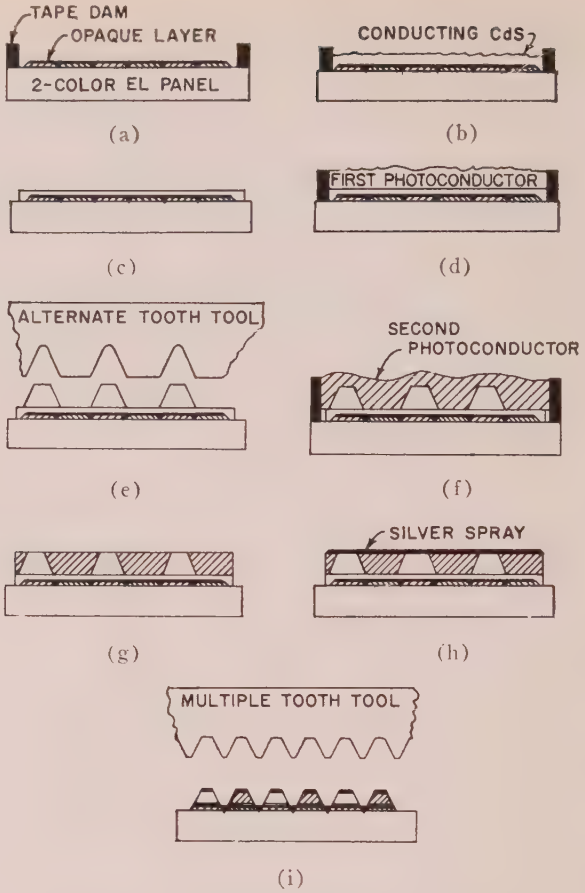

Fig. 4—Steps in construction of two-color photoconductive input layer.

under them can come loose [Fig. 3(c)], leaving alternate bands of blue phosphor and conducting-coating covered glass. A second spray, this time yellow phosphor, is applied and baked [Fig. 3(d)]. Then the whole surface is levelled to about one mil thickness and the strips made symmetrical by a second application of the brass lap, which cuts through the transparent conducting coating, thus electrically isolating each line from the adjacent ones [Fig. 3(e)]. A spray of opaque paint completes the output stage of the panel [Fig. 3(f)].

The intermediate current diffusing layer is then applied by casting [Fig. 4(a) and Fig 4(b)], using conducting CdS powder in diluted epoxy resin, Ciba Araldite, 1 part to 5 parts diacetone alcohol. After machining the conducting layer to about five mils thick [Fig. 4(c)], the first photoconductor, CdSe powder, is cast into place using the same binder [Fig. 4(d)]. When throughly cured, the layer is grooved, parallel to the reference edge, using an alternate tooth tool mounted on a hand-driven shaper so that ridges of photoconductor remain above the bands of yellow electroluminescent phosphor [Fig. 4(e)].

Then the second photoconductor, CdS, is applied also by casting, so that the grooves between the ridges are filled [Fig. 4(f)]. To prevent loss of sensitivity in the first photoconductor this second layer is made using a binder of 5 per cent polystyrene in pentachlorethane. After the drying of this layer, the surface is machined flat [Fig. 4(g)], and sprayed with conducting silver [Fig. 4(h)]. The next step is the final cutting of the ridges into two separate sets using a multiple tooth tool [Fig. 4(i)], and cutting down through the photoconductor and the conducting layer until the opaque layer is just exposed. In order to obtain separate electrical contact to the lines on the selenide and the lines on the sulfide, this last machining step is not carried across the panel: The tool instead is lowered into the panel and withdrawn in order to leave about $\frac{3}{16}$ inch of continuous silver at both ends. It is then a simple matter, using the alternate tooth tool, to disconnect one set at one end by cutting the silver at the top of the ridges and to disconnect the other set at the opposite end leaving two sets of interdigitated lines.

A photomicrograph of an actual light amplifier in cross section appears in Fig. 5. Below the photograph is a line drawing which identifies the various parts. It will be noted that each ridge of photoconductor has associated with it only a single strip of phosphor; the electrical crosstalk due to conduction from one ridge to an adjacent phosphor strip is virtually eliminated both because the current diffusing layer is nearly cut through, and because of the high field dependence of the current in this material.

Experimental Results

The significant measurements on the two-color light amplifier are essentially the same as those for a monochrome device except that separate results are obtained for each color output. Fig. 6 is a plot of output luminance (foot-lamberts) as a function of input illuminance (foot-candles) for a tungsten light source. A graph is shown for excitation of the cadmium sulfide

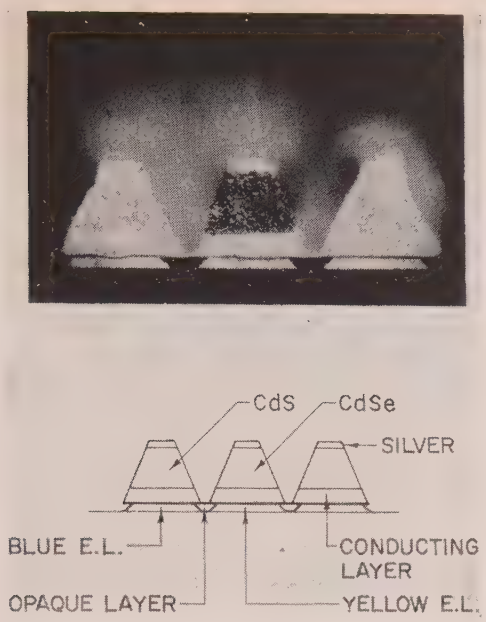

Fig. 5—Photomicrograph of cross section of completed two-color amplifier with line drawing for comparison.

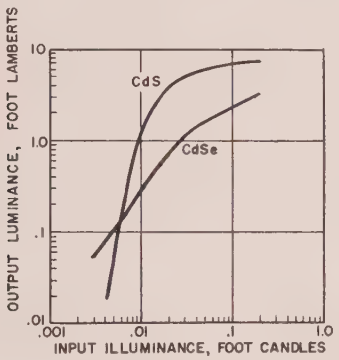

Fig. 6—Input-output curves for two-color light amplifier.

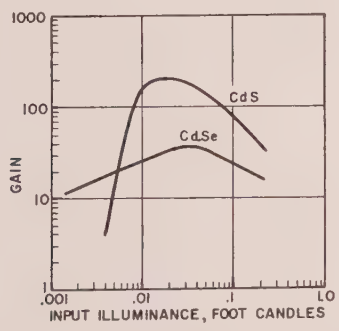

Fig. 7—Gain curves for two-color light amplifier.

ridges and also for excitation of the cadmium selenide ridges. In the case of the sulfide the output light is from the blue emitting phosphor strips only, while excitation of the selenide ridges causes emission from the yellow emitting phosphor only. Both curves show rather high gamma, but continuous tone images are possible in the region before the output begins to saturate.

Fig. 7 shows the tungsten light gain calculated from the previous curves. These show a higher gain for the sulfide photoconductor than for the selenide. In each case, however, if these gains are to be compared with monochrome amplifiers they should be multiplied by two since only half the area of the amplifier is active. The energy gains of the two inputs are difficult to measure because of the differences in spectral characteristics of the source, photoconductors and phosphors.

No quantitative measurements were made of the speed of response of the panel but for visible light input the response is very slow because of the use of cadmium sulfide powder, which has a rise and decay time of several seconds. For the infrared input on the cadmium selenide, the response is much faster and it is possible to view moving objects. The cadmium selenide has a rise and decay time of about 0.1 second.

In order to operate the two-color 6-inch×6-inch panel conveniently as an imaging device, a light-tight adaptor was made to hold the panel and to fit the back of a 4-inch×5-inch Speed Graphic camera. The adaptor incorporated the necessary electrical connections; with a special short focus lens on the camera, the panel could be operated with an image source, 24-inch×24-inch placed at a distance of six feet.

The two-color panel images a source of visible light (with no infrared) where the sulfide photoconductor is excited and images a source of near-infrared where the selenide is excited. Two light sources with these characteristics can be independently imaged in blue and yellow respectively. In order to observe the two-color output images on the panel, it is necessary to simulate a scene incorporating these two kinds of illumination. Such an image source is made by illumination of a scene reproduced in black paint on an opal glass twenty-four inches square. A visible source with no infrared is obtained by uniformly illuminating the whole opal glass image with a number of daylight fluorescent lamps. The invisible near infrared sources are obtained by filtering the output from tungsten lamps with Corning 2540 filters. Fig. 1 gives the spectral characteristics of the two sources and the response of the sulfide and selenide photoconductors. A photograph of the illuminated opal glass is shown in Fig. 8(a) as it appears to the eye. However, in the two dotted squares, one in a black area and one in a white area, there is also incorporated a filtered tungsten light source which is invisible to the eye.

The opal glass screen is imaged through a lens onto the photoconductive surface of the light amplifier. Alternating voltage at 420 cps is applied to the panel between the transparent conducting coating and each set of silver lines on the photoconductive ridges. Each voltage is adjusted separately on the sulfide and selenide ridges so that in the dark there is no blue or yellow output. In addition, further adjustment is made, if necessary, to eliminate any cross talk such as the blue lines emitting light when the selenide is excited or the yellow lines emitting light when the sulfide is excited. Suitable voltages on a particular amplifier are 500-volt rms for the sulfide and 300-volt rms for the selenide.

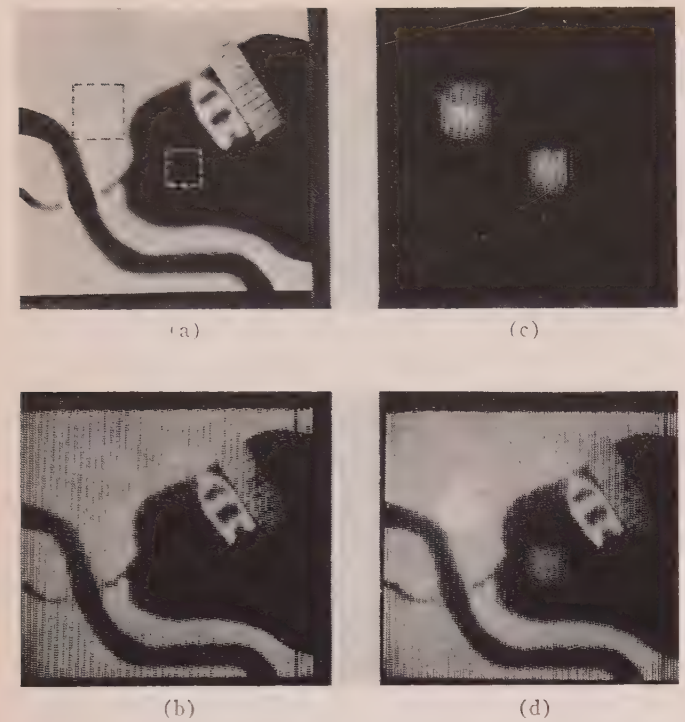

Fig. 8—(a) Opal glass image source, (b) light amplifier output with visible input only, (c) light amplifier output with infrared input only, (d) light amplifier output with both inputs.

Fig. 8(b) shows a photograph of the output for visible input only with a suitable voltage applied to each photoconductor. The image on the amplifier output is, of course, blue in color. Fig. 8(c) is a photograph of the panel output (yellow in color) when only the invisible infrared parts of the opal glass image are illuminated. Finally in Fig. 8(d) a photograph of the output image is shown with both visible and invisible inputs operating. The two superimposed output images are in blue and yellow respectively and where they overlap an approximate white is formed. On careful examination, it is possible to observe from the photographs that the yellow emitting lines are interleaved with the blue emitting lines. These photographs illustrate the ability of the panel to detect and separate simultaneous visible and invisible inputs incident on the panel.

Acknowledgment

The authors wish to thank D. W. Epstein and H. Johnson, of RCA, and R. K. H. Gebel of the Wright Air Development Division, for their continued interest in this paper. Much of the detailed constructional work was done by H. Ogawa. T. Dettre and G. Arleth deserve special credit for executing the difficult lapping and machining operations.

Construction of a Thermionic Energy Converter*

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Summary—**Some of the general design considerations for a cesium vapor-type thermionic energy converter are discussed. These include the questions of geometry and materials as well as cathode support and cathode lead design. The effect of the internal magnetic fields has been analyzed; it was found that there exists an upper optimum limit for the power handling capability per unit cell.**

The actual construction of a high power thermionic energy converter is described. This unit is of coaxial geometry having the anode surrounding the cathode and heat source.

Laboratory testing of this cell is described and includes cold vacuum testing as well as high power operation using an electrical heat source external to the cell. For a device having a molybdenum cathode at 2525°K, a power output of 270 watts was obtained. The efficiency was about 3 per cent, which is in fair agreement with design predictions.

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Introduction

THE thermionic energy converter (TEC) is a device for direct conversion of heat to electricity. The scientific principles of this device have been studied by several workers.[1–6] It is the object of this

[1] H. Moss, "Thermionic diodes as energy converters," *J. Electronics*, vol. 2, pp. 305–322; January, 1957.
[2] N. D. Morgulis and D. M. Marchuck, "Conversion of thermal energy into electrical energy by means of thermionic emission," *Ukrainian Phys. J.*, vol. 2, pp. 379–381; 1957.
[3] K. G. Hernqvist, M. Kanefsky, and F. H. Norman, "Thermionic energy converter," *RCA Rev.*, vol. 19, pp. 244–258; June, 1958.
[4] G. N. Hatsopoulos and J. Kaye, "Analysis and experimental results of a diode configuration thermoelectron engine," PROC. IRE, vol. 46, pp. 1574–1579; September, 1958.
[5] G. M. Grover, D. J. Roehling, E. W. Salmi, and R. W. Pidd, "Properties of a thermoelectric cell," *J. Appl. Phys.*, vol. 29, pp. 1611–1612; November, 1958.
[6] V. C. Wilson, "Conversion of heat to electricity by thermionic emission," *J. Appl. Phys.*, vol. 30, pp. 475–481; April, 1959.

paper to analyze some of the general construction principles for a cesium-vapor type TEC, and to describe the actual construction and testing of a cell suitable for a practical heat source. Application of this converter to a solid propellant rocket engine is described elsewhere.[7]

The TEC bears great resemblance to the electron tube in its construction. Techniques known in the fabrication of electron tubes, such as metal-to-ceramic seals, can be directly applied to the construction of the TEC. The TEC differs from conventional electron tubes primarily in the way heat is supplied to the cathode. Specifically, in the case when the heat source is external to the converter, the hot cathode and its support will have to be a part of the tube envelope. Thus, one of the main construction problems becomes the support of the hot cathode.

The cell geometry considered in this paper is shown in Fig. 1. The cell has a coaxial structure in which the

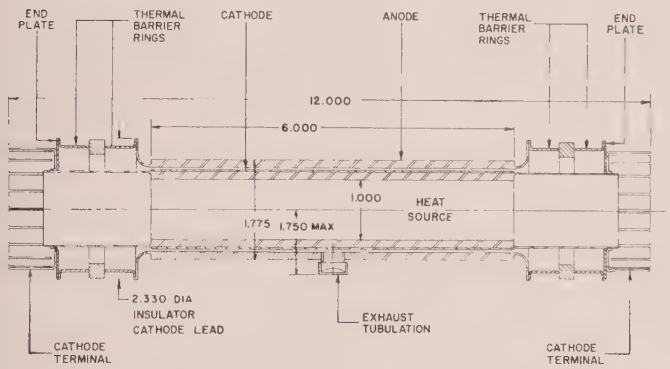

Fig. 1—Thermionic energy converter of cylindrical geometry.

anode surrounds the cathode and the heat source. In this structure the anode acts as a heat shield around the cathode and the conversion efficiency[3] may be written approximately

$$\eta_0 = \frac{(\phi_C - \phi_A)I}{\phi_C I + P_{\text{rad}}}, \tag{1}$$

where

ϕ_C = cathode work function,
ϕ_A = anode work function,
I = current,
P_{rad} = radiation loss form the cathode,

and the radiation losses become

$$P_{\text{rad}} = \sigma T_C^4 \left(\frac{1}{\epsilon_C} + \frac{1}{\epsilon_A} - 1\right)^{-1} A_C. \tag{2}$$

Here

T_C = cathode temperature in °K,
σ = Stefan-Boltzmann constant,
ϵ_C = thermal emissivity of the cathode,
ϵ_A = thermal emissivity of the anode,
A_C = cathode area.

The Cathode Lead Problem

The lead connecting the hot cathode to the colder cathode output terminal of the converter serves three purposes:

1) It supports the hot cathode
2) It provides thermal isolation for the hot cathode
3) It conducts the current from the cathode to the output terminal.

The first and third purposes require a cathode lead which is short and has a large cross section for structural rigidity and for minimization of the voltage drop in the lead. The second purpose, however, calls for a long and thin lead to minimize heat conduction loss from the cathode. Thus, a compromise must be made between these requirements.

If the cathode lead losses are included in the efficiency relation (1), the efficiency becomes[8]

$$\eta \cong \frac{(\phi_C - \phi_A)I - \rho \frac{L}{A} I^2}{\phi_C I + P_{\text{rad}} + \kappa \Delta T \frac{A}{L} - \frac{1}{2}\rho \frac{L}{A} I^2}, \tag{3}$$

where the properties of the lead are

ρ = resistivity,
L = length,
A = cross-sectional area,
κ = heat conductivity,

and

ΔT = temperature drop along lead, or approximately cathode temperature minus ambient temperature.

Since the voltage drop in the lead is

$$V = \rho \frac{L}{A} I, \tag{4}$$

(3) becomes

$$\eta = \frac{\phi_C - \phi_A - V}{\phi_C + \frac{P_{\text{rad}}}{I} + \frac{\kappa \rho \Delta T}{V} - \frac{V}{2}}. \tag{5}$$

[7] M. Yaffee, "Unit converts waste rocket heat to power," *Aviation Week*, vol. 71, pp. 92–101; November, 1959.

[8] J. M. Houston, "Theoretical efficiency of the thermionic energy converter," *J. Appl. Phys.*, vol. 30, pp. 486–487; April, 1959.

For optimum lead design

$$\left(\frac{\partial \eta}{\partial V} = 0\right)$$

relations for V and η are obtained which for most cases of practical interest reduce to

$$V \cong \sqrt{\kappa \rho \Delta T \eta_0} \tag{6}$$

and

$$\eta \cong \eta_0.$$

Eqs. (4) and (6) yield the lead design factor

$$\frac{L}{A} = \frac{1}{I}\sqrt{\frac{\kappa \Delta T \eta_0}{\rho}}. \tag{7}$$

Fortunately, the relation between η and V [see (5)] has a relatively broad maximum. This implies that considerable deviations from (7) are allowed without seriously decreasing the efficiency.

When the lead must support the cathode structural rigidity, *i.e.*, a short and thick lead, is called for. This requirement demands a low L/A value. Since the quantities under the square root sign of (7) are more or less predetermined by the limited choice of materials and specified device performance, a low L/A value calls for a large value for I, the current handling capability per unit cell. Other considerations, however, set an upper limit on I. For practical reasons it is desirable to operate at reasonably high value for the load impedance[3]

$$\frac{\phi_C - \phi_A}{I}.$$

Secondly, an upper limit on I is set by the effect of the internal magnetic fields generated by the currents flowing in the device. The effect of this magnetic field is considered below.

Effect of Internal Magnetic Fields

Consider a cylindrical geometry as shown in Fig. 2. The output current flows along the cathode towards both ends of the cathode. Thus, the axial current is zero at the center of the cathode and a maximum at the ends. A magnetic field is generated by this axial current with circular field lines enclosing the cathode. Thus, this magnetic field is perpendicular to the electron flow between cathode and anode and tends to impede the flow of electrons. The electrons, which travel only with thermal velocities in the interelectrode space, are caused to describe circular orbits, and some of the slow ones never reach the anode.

To determine the effect of the magnetic field on the current, consider a part of the axial cross-section of the device as shown in Fig. 3. The cathode and the anode are separated by distance a, and the strength of the magnetic field is B. The radius of the electron orbit is

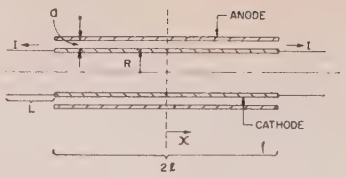

Fig. 2—Schematic diagram showing cylindrical converter.

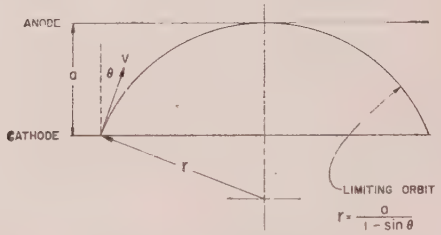

Fig. 3—Schematic drawing illustrating effect of internal magnetic field.

$r = mv/eB$, where v = electron velocity and m/e is the mass-to-charge ratio for the electron. For electrons emitted in the direction θ, only those having velocities larger than

$$v_{crit} = \frac{eBa}{m(1 - \sin\theta)}$$

will reach the anode.

The velocity distribution of emitted electrons can be written in terms of the current density

$$\frac{dJ}{J_0} = K \exp\left(-\frac{1}{2}\frac{mv^2}{kT_C}\right) v^3 dv \cos\theta d\theta$$

where J_0 is the emission current density, K is a normalizing constant, and $k = 1.371 \times 10^{-23}$ joule/°K is the Boltzmann constant. K is such as to make the integral

$$K \int_{-\pi/2}^{+\pi/2} \cos\theta d\theta \int_0^{\infty} \exp\left(-\frac{1}{2}\frac{mv^2}{kT_C}\right) v^3 dv = 1.$$

For the case of a magnetic field, the integration limits for v are v_{crit} and ∞. The integration gives the relation between actual current density and magnetic field:

$$J = J_0\left(\exp(-b^2) - \frac{\sqrt{\pi}}{2} b[1 - \mathrm{Erf}(b)]\right) \tag{8}$$

where

$$b = eBa(8mkT_C)^{-1/2}.$$

For a cylindrical geometry, the magnetic field at any point is given by the axial current flowing along the cathode. Using the notation shown in Fig. 2, the field is given by

$$B = \mu_0 \int_0^x J dx \tag{9}$$

where $B(x)$ is the magnetic field at point x, and μ_0 is the permeability of vacuum. Eqs. (8) and (9) uniquely define the problem of finding the actual current density as a function of x.

The total current from half of the cathode is

$$I = \int_0^l 2\pi RJdx, \tag{10}$$

where R is the cathode radius and $2l$ is the cathode length. Eqs. (8), (9), and (10) may be written in the following normalized forms:

$$b = \frac{e\mu_0}{2\pi}(8mk)^{-1/2}\frac{I_0 a}{R\sqrt{T_C}}\int_0^y F(b)dy,$$

$$\frac{I}{I_0} = \int_0^1 F(b)dy,$$

$$F(b) = \exp(-b^2) - \frac{\sqrt{\pi}}{2}b[1 - \mathrm{Erf}(b)]. \tag{11}$$

Here I_0 is the current in the absence of a magnetic field, or

$$I_0 = 2\pi RlJ_0. \tag{12}$$

The results of solving the simultaneous equations (11) with a computer are shown in Fig. 4. Here I/I_0 is plotted as a function of the parameter $I_0a/R\sqrt{T_C}$, where I_0 is expressed in amperes, T_C in °K, and a and R are lengths expressed in equal units. Also shown in Fig. 4 are corresponding results for a geometry in which the cathode and the anode are plane, parallel, circular disks of radii R, separated by distance a. In the computation of the disk structure, it is assumed that the current is drawn out uniformly along the cathode edge, and that the anode current is flowing in such a way as to maintain circular symmetry for the current flow pattern. To a first approximation the actual conversion efficiency is $\eta_0 I/I_0$. Thus it is seen from Fig. 4 that in order not to reduce the efficiency by more than 10 per cent, the parameter $I_0a/R\sqrt{T_C}$ must not exceed about 0.075. For given values of J_0, l/R, a, and T_C, this requirement sets a practical upper limit on the current handling capability per unit cell.

Choice of Cathode and Anode Materials

The thermionic converter to be described here is of the cesium vapor type.[2,3,5,6] Cesium ions for space-charge neutralization are generated by reasonance ionization at the cathode. Resonance ionization requires that the cathode, or at least part thereof, have a work function higher than the ionization potential of cesium (~3.9 volts). Of the refractory metals, tantalum or molybdenum are preferable for their high emission current density. For instance, at a temperature of 2500°K,

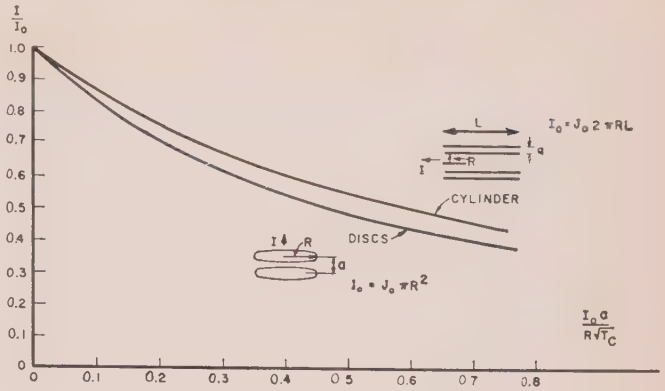

Fig. 4—Diagram showing effect of internal magnetic field. I_0 = current density × cathode area, I = actual delivered current, a = separation between cathode and anode, R = cathode radius.

the emission current density[9] for tantalum is about 3.5 A/cm² and for molybdenum about 0.8 A/cm². The life of the device is primarily determined by the evaporation rate[10] of the cathode. At 2500°K about 0.3 mm of tantalum and 2 mm of molybdenum are evaporated off in one year.

The cesium also serves another purpose in this tube. It covers the anode to give an anode work function of about 1.8 volts.[3] Any metal which is not attacked by the cesium[11] may then serve as anode material.

Specific Device Parameters

In the construction of the thermionic energy converter described in this paper, the nature of the heat source was considered as a fixed design criterion. The heat source is such as to permit a cylindrical cathode surface of 1.64 cm radius and 6 inch length. The cathode emission area is then 150 cm². Furthermore, a cathode temperature at 2500°K is assumed. Two cathode materials are considered, namely molybdenum and tantalum. The physical properties[9,12] for these materials used in the computations are shown in Table I.

The anode surface is cesium on nickel-plated copper, yielding a work function of 1.8 volts and a thermal emissivity[13] of about $\frac{1}{3}$. The anode is assumed to be the coldest part of the device with a temperature in the range of 150–250°C. This temperature range ensures a cesium vapor pressure of 10^{-2} to 10^{-1} mm of Hg, which is sufficient for space charge neutralization for the electron current densities of interest.[3] A cathode-to-anode spacing of 1 millimeter is assumed. The spacing is small enough to ensure a negligible scattering of electrons in the interelectrode space,[3] yet large enough to make fabrication easy and to minimize the danger of cathode-to-

[9] W. Espe and M. Knoll, "Werkstoffkunde der Hochvakuumtechnik," Julius Springer Verlag, Berlin, Germany, p. 258; 1936.
[10] S. Dushman, "Scientific Foundations of Vacuum Technique," John Wiley and Sons, Inc., New York, N. Y., p. 745; 1949.
[11] A. W. Hull, E. E. Burger, and R. E. Turrentine, "The cesium-vapor rectifier," *GE Rev.*, vol. 54, pp. 16–22; August, 1951.
[12] C. J. Smithells, "Metals Reference Book," Butterworth's Scientific Publications, London, England, vol. 2, pp. 666 and 638; 1955.
[13] H. B. Ives and H. B. Briggs, "Optical constants of rubidium and cesium," *J. Opt. Soc. Am.*, vol. 27, pp. 395–400; November, 1937.

TABLE I

DEVICE PARAMETERS FOR CESIUM VAPOR TYPE THERMIONIC ENERGY CONVERTER

Cathode	Tantalum Cathode	Molybdenum Cathode
Temperature 2500°K		
Cathode radius 1.64 cm		
Cathode length 6 inches		
Cathode area 150 cm²		
Emission current density	3.5 A/cm²	0.8 A/cm²
Cathode current $2I_0$	520 A	120 A
Work function	~4 V	~4 V
Thermal emissivity	1/3	1/3
Magnetic Field Effect		
$I_0a/R\sqrt{T_C}$	0.32	0.073
I/I_0	0.66	0.9
Cathode Lead		
Material molybdenum		
Length 1.5 inch		
Thickness	0.014 inch	0.007 inch
Device Performance		
Current	350 A	108 A
Voltage	~2 v	~2 v
Power out	700 w	216 w
Power input	8050 w	7515 w
Efficiency	8.7 per cent	2.9 per cent

anode short circuits due to thermal expansion and warping of the cathode. See Table I.

The calculated effect of the internal magnetic field is shown in Table I. It is seen that the magnetic field results in an appreciable reduction of current (and thus of efficiency) for the case of a tantalum cathode. Thus, for optimum efficiency a multiplicity of smaller converters is required. The expected performance for a single device is shown in Table I.

The cathode lead is assumed to be a cylindrical thin-wall extension of the molybdenum tube. For a 1.5-inch-long lead, the wall thickness calculated from (7) is shown in Table I for the two cases considered. Here it is assumed[12] that $\kappa = 0.84$ watt/cm$\times$°C, $\Delta T = 2000$°C, and $\rho = 46$ μohm$\times$cm. Because of the relative insensitivity of the efficiency to a deviation from optimum lead design, a wall thickness of 0.013 inch was chosen in the actual converter as a compromise to assure structural rigidity for the lead.

FABRICATION OF PARTS FOR THE CONVERTER

The device (see Fig. 1) consists of an outer shell assembly, an inner cathode tube, two end plates connecting the outer shell to the inner tube, and two cathode terminals attached to the end plates. The inner tube, the cathode, is machined for a solid bar of pure molybdenum, made by sintering and hot swaging in controlled atmospheres. The end sections have walls 0.013 inch thick. Their purpose is 1) to retard heat loss from the cathode area, 2) to conduct the current from the cathode to the terminals, 3) to support the cathode, and 4) to form a part of the tube envelope.

In devices using tantalum as the emitting surface, a 0.020-inch-thick tantalum tube is shrunk around the molybdenum tube, as shown in Fig. 1. The end plates are made of 0.020-inch-thick molybdenum. The diaphragm-like shape permits a certain amount of deflection and conforms to the general need for thermal isolation. The thermal barrier rings are made from 0.030-inch-thick kovar, chosen both for its low heat conductivity and its coefficient of thermal expansion, which matches closely that of the ceramic insulator.

The ceramic insulator is a high alumina ceramic, metallized on both sides to permit brazing to kovar and copper, respectively.

The anode is made of OFHC copper, with flexible stress isolation sections at each end.

ASSEMBLY OF THE CONVERTER

The copper terminals and the molybdenum end plates are first made into subassemblies by brazing with Nicoro, a Ni-Cu-Au alloy, in an oxidized stainless steel jig which aligns the parts.

A special jig was designed in which one of the terminal subassemblies together with the other parts forming the outer shell are brazed into one single unit.

This jig, also of oxidized stainless steel, performs these functions:

1) It aligns all parts concentrically.
2) It supports the thin sections of the anode, prevents them from collapsing, and maintains the over-all length of the device within close limits.
3) It shields the entire assembly, and particularly the ceramics, against direct radiation in the furnace while promoting even heating and cooling.
4) It provides adequate venting for the passage of hydrogen which can readily replace all air inside the unit.
5) It places the correct weight on the vertically stacked assembly for optimum brazing conditions.

This brazing operation is done at 820°C, using a silver-copper eutectic solder.

The next step is to insert the cathode tube. At the top, it fits snugly into the hole of the end plate. At the bottom, a concentric spacer aligns the tube to the inside diameter of the anode. Thus assembled, it is placed vertically inside the "bell" of an RF furnace with hydrogen atmosphere. A specially designed induction coil permits the joint to be heated very quickly to the melting point of Nicoro without melting the nearby joints of Nicoro and silver solder, respectively.

The unit is now turned upside down, the second subassembly is slipped over the cathode tube, and the moly end plate is brazed to the kovar ring with Nicoro in the RF furnace. The last braze, namely cathode tube to end plate, follows, again using Nicoro and RF. The unit is inspected and leak checked.

The device is then exhausted and baked at 400°C for two hours. During bake-out the tube envelope is protected by an inert gas. The cathode is heated to 1200°C

using a glowbar. Finally, cesium is introduced and the device sealed off. Fig. 5 is a photograph of the finished converter. Fig. 6 shows an exploded view of the device.

Testing of the Converter

A scheme for cold vacuum testing of the finished converter has been devised. Usually, cesium vapor deposits on the insulators, which causes a cathode-to-anode short circuit. By placing the anode on a hot plate (150°C) and cooling the cathode, the cesium deposits on the cathode and opens up the cathode-to-anode short circuit. This indicates the presence of free cesium in the tube, and thus a very low oxygen, hydrogen, and nitrogen pressure (cesium "getters" these gases). If a voltage of 1000 volts can be applied between cathode and anode without leading to breakdown, a sufficiently low pressure of other gases is assured. In a tube which originally shows no cathode-to-anode short, another test of the presence of free cesium can be made by heating the whole tube to 150–200°C in an oven. An applied voltage of 1000 volts will lead to glow discharge breakdown in the cesium vapor.

For the performance test of the device, special test equipment was designed. The cathode was heated by radiation from a cylindrical, self-supporting tantalum heater capable of dissipating 10 kw at approximately 30 volts. Provision was made for electron bombardment of the cathode, using thermionic emission from this heater as a supplement to the radiant heating. The anode cooling was achieved by means of a removable cooling course clamped to the anode. The design was of such a nature as to permit cooling by water, steam, or a gas such as compressed air, in order to provide temperature control over a wide range of power and temperature levels. Separate cooling of the end seals and the output connectors was provided. Protection of the converter, heater, and supporting mechanism from atmospheric attack was achieved by enclosing the entire structure in a high vacuum bell jar. This also provided the proper conditions for electron bombardment of the cathode.

A fixed electrical load of 0.01 ohm was constructed. The circuitry was so arranged that voltage and current could be read under both loaded and unloaded conditions.

The measurement of cathode and heater temperatures was made by means of an optical pyrometer. The temperatures of the ceramic seals and of the anode were continuously monitored by means of suitable thermocouples whose leads were brought out of the bell jar through vacuum seals. The entire bell jar was supported on the manifold of a standard exhaust system employing an oil diffusion pump, as shown in Fig. 7.

The bell jar was pumped to a pressure in the order of 10^{-5} mm Hg. The heater temperature was raised slowly in order to permit gradual outgassing of both the converter and the bell jar. At a cathode temperature in the order of 2100°K the first signs of life would appear in the form of a small open-circuit voltage across the converter. Upon raising the cathode temperature to 2500°K,

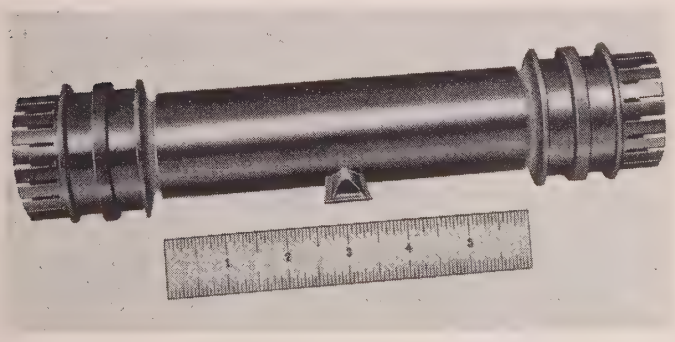

Fig. 5—Photograph showing the completed thermionic converter.

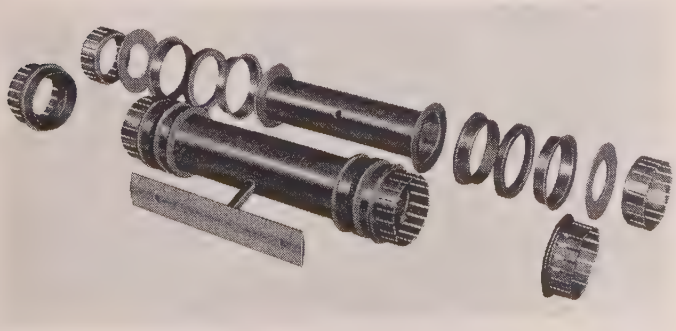

Fig 6—An exploded view of the device.

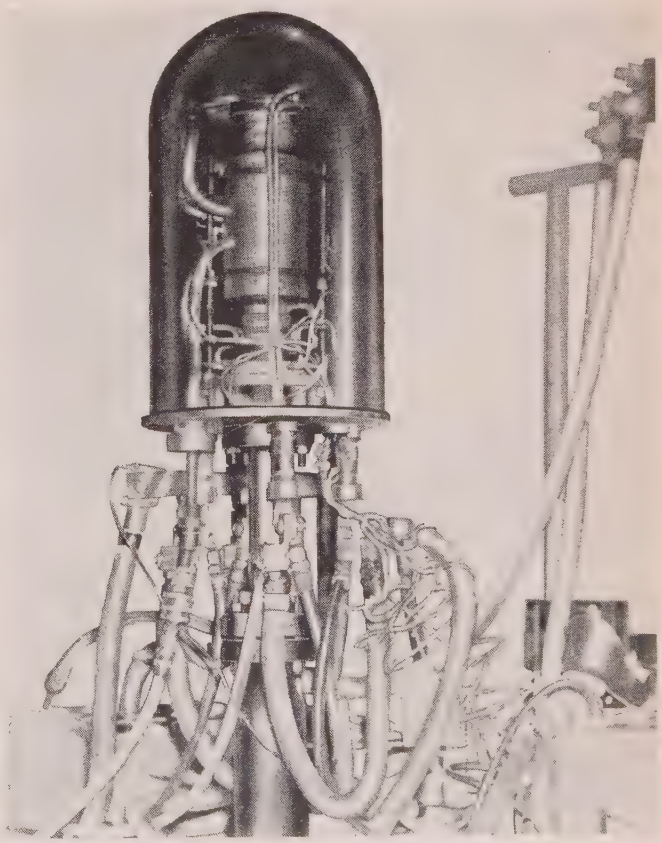

Fig. 7—Photograph of laboratory test apparatus.

substantial output was obtained. During this test period, the anode temperature was maintained in a range between 150 and 250°C.

Results of tests on a converter having a molybdenum cathode are shown in Table II, which also shows the results corrected for optimum load as well as the calculated values for current[9] and efficiency. Considering the fact that a substantial fraction of the heat was lost by conduction through the heater leads and radiation at the unshielded ends of the device, these results are consistent with the predicted ones.

TABLE II
TEST RESULTS FOR TEC HAVING MOLYBDENUM CATHODE

	Measured	Corrected for Optimum Load	Calculated
Cathode Temperature	2525°K		
Anode Temperature	150–250°C		
Current	164 A		
Current Density	1.1 A/cm²		1.1 A/cm²
Load Impedance	0.01 ohm		
Voltage	1.64 v	~2 v	
Power Output	269 w	328 w	328 w
Power input Heater Bombardment	 3200 w 6900 w		7596 w
Total	10100 w		
Efficiency	2.7 per cent	3.25 per cent	4.1 per cent

Conclusions

In conclusion, the successful construction of a high-temperature cesium vapor-type thermionic energy converter, which has an external heat source and which can be operated in vacuum or in an inert atmosphere, has been demonstrated.

All the techniques used in the construction of the cell are commonly used in the manufacture of electron tubes and can be readily applied to quantity production.

Nanosecond Response and Attenuation Characteristics of a Superconductive Coaxial Line*

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Summary—**A miniature superconducting coaxial line is described which transmits nanosecond (nsec) pulses without any measurable change in rise time. Data in regard to the temperature dependence of the bandwidth over the range 293°K to 4.2°K are presented. The investigation is confined to ten and hundred foot lengths of cable.**

Introduction

THE TRANSMISSION of a nanosecond (10^{-9} second) duration pulse over a relatively long transmission path is conveniently accomplished by a coaxial transmission line; the bandwidth of the line is primarily limited by its inherent skin effect attenuation.[1] Physically realizable coaxial lines are not uniform lines because connectors, dielectric supports, and random geometrical imperfections introduce nonuniformities which in turn produce reflective losses and may also produce higher order mode transmission of microwave frequencies. Skin effect losses for a given characteristic impedance may be reduced to any desired nonzero magnitude by increasing the respective dimensions of the coaxial line; however, mode conversion effects, if they are present, will occur at much lower frequencies.[1] Nanosecond systems by their very nature are inherently small physically and are not too compatible with large diameter coaxial lines. In nsec pulse oscillography it is usual practice to employ relatively large semirigid 50-ohm coaxial lines ($\frac{1}{2}$ inch to $\frac{7}{8}$ inch outer diameter) to provide signal delays of the order of 100 nsec in the vertical signal channel. Such lines possess an attenuation which is less than 10 db/100 feet at 4000 Mc.

This paper describes a miniature coaxial line (50 ohms, 0.091 inch diameter) that possesses a very low attenuation by virtue of the superconductivity phenomenon which occurs at low temperatures. The miniature line was constructed for use as a relatively lossless signal delay line for nsec pulse oscillography.

* Received by the IRE, May 18, 1960; revised manuscript received, August 29, 1960.

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[1] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y.; May, 1956.

The Miniature Line

The miniature line was fabricated in a total length of 200 feet; its construction is as follows:

1) Inner conductor, niobium, 0.010 inch diameter;
2) Outer conductor, lead, 0.034 inch inner diameter, 0.091 inch outer diameter;
3) Dielectric, teflon.

Electrically, the line has a characteristic impedance of 50 ohms and a capacity of 29.1 picofarads per foot. Mode conversion occurs for frequencies on the order of 100 kMc. The line superconducts for temperatures less than 7.22°K, the critical temperature for Pb; the critical temperature for Nb is 8.7°K. The measured conductivities of the conductors at 293°K. (room temperature) are 6.14×10^6 mhos/meter for the Nb center conductor and 4.82×10^6 mhos/meter for the Pb outer conductor. Fig. 1 shows a photograph of the line and the associated transition to a type-*N* coaxial connector (note the boot match shown for size comparison).

The high-frequency attenuation of the line at 293°K was predicted by calculating the skin effect and dielectric losses. For liquid nitrogen temperatures (77.3°K) the dielectric losses were neglected, and the skin effect loss was based upon 77.3°K conductivities which were determined by direct measurement in a liquid N bath. No predictions were made as to the low temperature (4.2°K) attenuation of the line due to the fact that the superconductive losses would be very small and would not present any measurable value in the contemplated instrumentation system. This may be seen by referring to Fig. 2, an approximation to Gruneisen's curve.[2] The low temperature resistivity varies approximately as T^5, while for high temperatures it varies approximately as T. For the specific metal samples used in the line, Gruneisen's curve in conjunction with the two resistivity measurements (in the linear region of the curve) yields the resistivity constants[3-5]

$$\text{Nb: } \theta = 26.5; \qquad R_\theta = 1.3 \times 10^{-8} \text{ ohm meters}$$

$$\text{Pb: } \theta = 70°\text{K}; \qquad R_\theta = 4.49 \times 10^{-8} \text{ ohm meters.}$$

When the temperature is lowered, the normal resistivity (not the superconductive resistivity) approaches the residual resistivity (resistivity at 0°K). Fig. 2 shows that the normal conductivity at low temperatures be-

Fig. 1—The miniature line and its associated transition to a type-*N* connector.

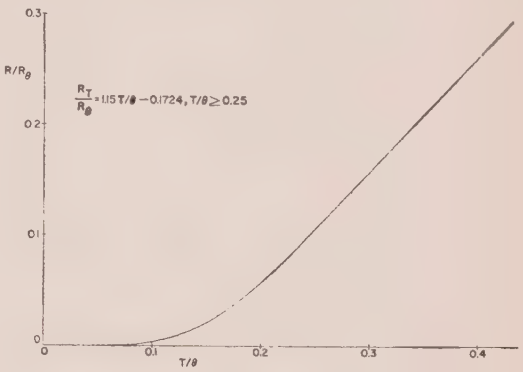


Fig. 2—Gruneisen's curve, $\theta \equiv$ Debye temperature, $R_\theta \equiv$ Debye temperature resistivity.

comes very large. If at low temperatures the conductivity is increased by a factor of 10^4, the line attenuation at any RF frequency would be reduced by a factor of 10^2; this in itself would be a remarkable reduction in attenuation which could not be fully detected by the contemplated instrumentation system. When superconductivity occurs, the conductivity (Pb) may increase by a factor of 10^{17} over the 293°K conductivity.[3]

The Experimental System

The experimental system is shown in block diagram form in Fig. 3(a) and 3(b). The insertion switch, which may be seen in Fig. 4, provides a convenient method for rapid insertion switching in the rigid coaxial system. Rigid coaxial lines are used to minimize the loss and instability in the over-all transmission system. The pulse transmission measurements were accomplished on a Project JAYHAWK traveling-wave synchroscope that employed a DuMont type KR 1524 P11M TWCRT.[6] The deflection response of the indicator is 6 db down at 2.1 kMc. The 100 feet of miniature line when wound on a 2 inch diameter form constitutes a cylindrical package of $3\frac{1}{4}$ inches in diameter and 5 inches long. The low-temperature system consists of two concentric dewars, the He flask having an internal radius of $2\frac{7}{8}$ inches and a depth of $15\frac{1}{4}$ inches; the N flask has an internal radius

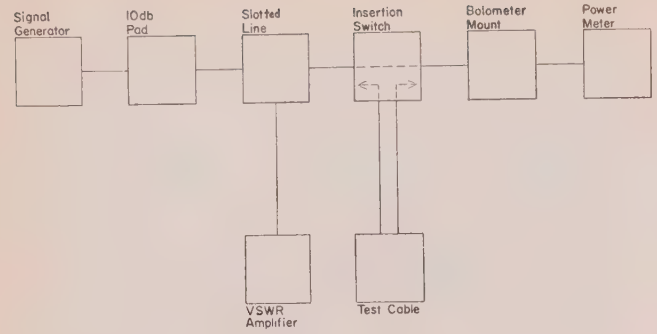
[2] C. Kittel, "Introduction to Solid State Physics," John Wiley and Sons, Inc., New York, N. Y.; August, 1955.

[3] The Pb values are of the same order as the usual values cited in the literature. On the other hand, the Nb Debye temperature is much smaller than the usual values (26.5°K compared to 252°K). This is not unusual; even the exact Gruneisen curve does not always agree with exeriment when the metal concerned is one of the transition metals. Fig. 2 is not intended to be quantitative, only qualitative; it is included at this point mainly to demonstrate qualitatively the variation of conductivity with temperature.

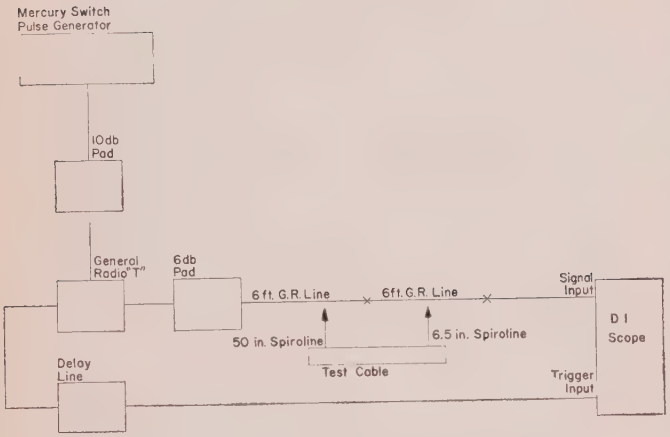
[4] "American Institute of Physics Handbook," McGraw-Hill Book Co., Inc., New York, N. Y.; 1957. Also, R. B. Scott, "Cryogenic Engineering," D. Van Nostrand Co., Inc., New York, N. Y.; 1959.

[5] N. F. Mott and H. Jones, "The Theory of the Properties of Metal and Alloys," Clarendon Press, Oxford, Eng.; 1936. (Reprinted by Dover Publications, Inc., New York, N. Y.; 1958.)

[6] Project JAYHAWK Progress Report, Contract No. DA 49-170-sc-2618, Univ. of Kansas Res. Foundation, Lawrence; April 12, 1959 to July 11, 1959.



(a)



(b)

Fig. 3—(a) Attenuation instrumentation. (b) Pulse response instrumentation.

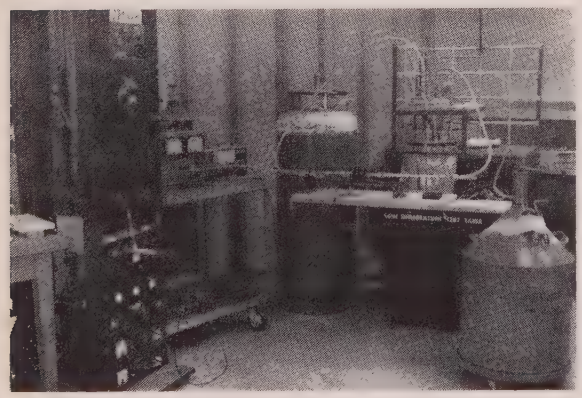


Fig. 4—Laboratory arrangement for attenuation measurements.

of $4\frac{1}{2}$ inches and a depth of $18\frac{1}{2}$ inches. The low-temperature operation of the system was at 4.2°K and He pumping was not required.

Experimental Results

The pulse response measurements employed an input pulse having the following characteristics:

1) 50 per cent pulse width 1.7 nsec.
2) 10 per cent–90 per cent rise time 0.4 nsec.
3) Amplitude 16.7 volts.

The pulse measurements were run on a 10 foot and a 100 foot length of line in order to demonstrate clearly the behavior of the line. The shorter length was necessary to obtain measurable responses when the line was operating at temperatures equal to or greater than 77.3°K [Fig. 5(a) and 5(b)]. Fig. 5(c) through 5(e) presents a graphic picture of the increase in bandwidth as the temperature moves from 293°K to 77.3°K to 4.2°K. Fig. 5(f) contains a superposition of the input pulse and the output pulse from the 100-foot line at 4.2°K. Notice that the rise time of the pulse is unaltered after passing through 100 feet of line, and that the upper portion of the output pulse has been reduced initially in amplitude by a reflection. The attenuation measurements demonstrate the presence of reflections.

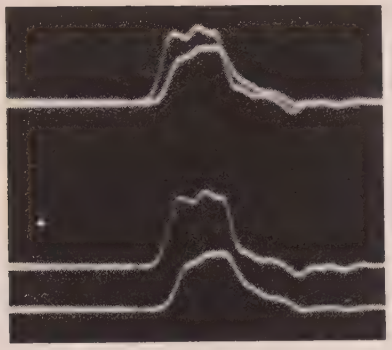
The attenuation measurements were also concerned with two lengths of line, 10 and 100 feet long. For temperatures equal to or greater than 77.3°K, the 10 foot length was employed; the results were multiplied by 10 and plotted in db/100 feet. In Fig. 6, *A* and *B* are the 293°K calculated and measured attenuation, respectively, and *C* is included the show the 293°K attenuation of an all-copper coaxial line having the identical dimensions of the miniature line. The attenuation of the copper line is less than the miniature line at 293°K, but is useless at 293°K for nsec pulse transmission. In Fig. 7, *A* and *B* show the measured and predicted 77.3°K attenuation.

In Fig. 8, *B* is the 4.2°K measured attenuation, while *A* is also included for comparison. This curve is the attenuation curve at room temperature of 100 feet of $\frac{1}{2}$-inch Spiroline rigid coaxial line. The 4.2°K attenuation curve has been adjusted in that the known rigid line losses and the input reflection loss have been subtracted from the over-all attenuation. Experimental error is in evidence which accounts for the negative attenuation. The variations in the 4.2°K attenuation are due to the multiple reflections that are present in the over-all system; however, even with the reflective losses the line outperforms the large 293°K Spiroline coaxial line. The reflection losses of the over-all system including all transmission paths are clearly placed in evidence when the miniature line is in the superconductive state (4.2°K).

When the majority of the attenuation disappears in the transmission line system, all of the nonuniformities of the system become evident and are exhibited in the undamped reflections produced by the system. Some of the discontinuities of the various components in the given system can be measured separately, and their effect on the over-all transmission may be calculated.

Regarding the discontinuities of the superconductive line and its associated connectors, it is found that the very low operating temperature causes geometrical changes (contractions) in the line itself and significantly disturbs the relative positions of the line conductors in the connector constant impedance matching transitions. This, in turn, produces a relatively large frequency-dependent VSWR that is not experimentally repeatable due to the present design of the transitions.

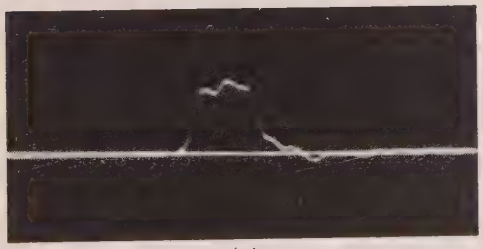
(a)



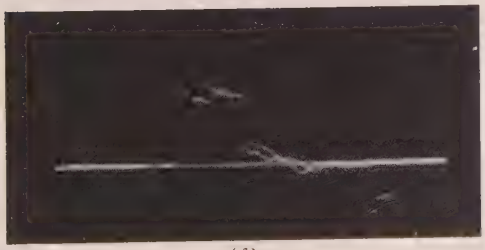
(b)

(c)

(d)



(e)



(f)

Fig. 5—(a) Upper trace: input pulse. Lower trace: output pulse, 10 feet, 293°K. (b) Upper trace: input pulse. Lower trace: output pulse, 10 feet, 77.3°K. (c) Output pulse, 100 feet, 293°K. (d) Output pulse, 100 feet, 77.3°K. (e) Output pulse, 100 feet, 4.2°K. (f) Superposition of input and output pulses, 100 feet, 4.2°K; input pulse shifted to left.

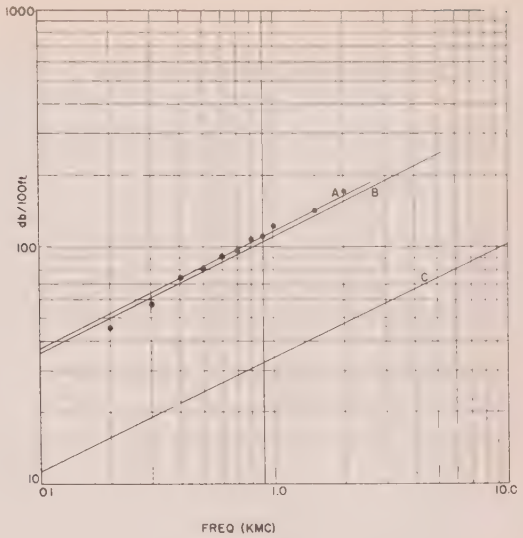


Fig. 6—Attenuation at 293°K.

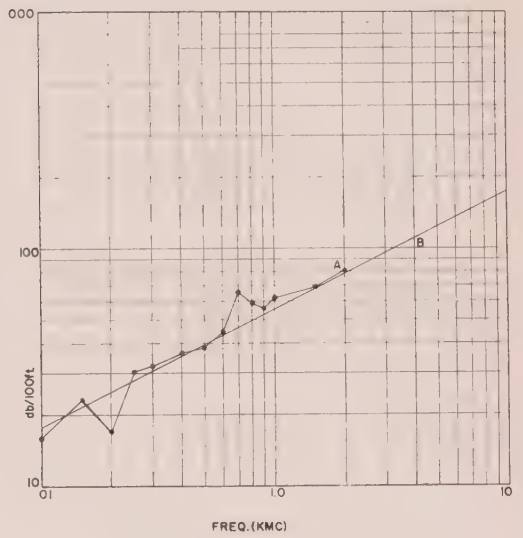


Fig. 7—Attenuation at 77.3°K.

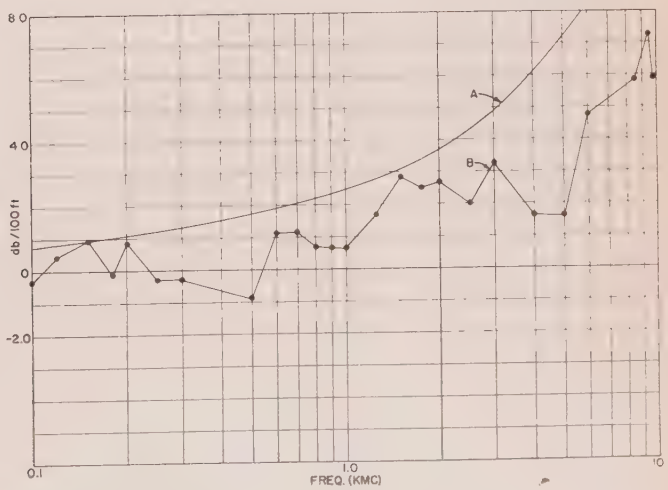


Fig. 8—$A = 100$ feet $\frac{1}{2}$-inch Spiroline cable (manufacturer's data) 293°K; $B = 100$ feet miniature line, 4.2°K.

Additional Remarks

The 4.2°K attenuation measurements in Fig. 8 are not to be interpreted as being indicative of the high-frequency attenuation to be found in superconductors. The majority of the resistive attenuation is due to the two short lengths of the miniature line that are not superconducting along their entire lengths. These two lengths are the input and output portions of the line that are not submerged in liquid He and are attached to the connectors shown in Fig. 1. The connectors are panel mounted in the base of a liquid N bucket; hence, the miniature line's extremities are held at 77.3°K.

In addition to nonuniform thermal contractions, the line attenuation is dependent upon the electrical characteristics of the particular Nb, Pb and teflon samples employed in the construction of the line. The fabrication techniques of the individual materials and the line itself introduce into the respective materials various stresses and strains which, in turn, alter their electrical characteristics. The Pb outer conductor was fabricated about the teflon coated Nb center conductor by a pressure-welding technique. The Pb was chemically pure, and subsequent measurements on the completed line showed that the conductivity was 4.82×10^6 mhos/meter at 293°K and 20.17×10^6 mhos/meter at 77.3°K. Using these values and tabulated values of Gruneisen's function[5] the Debye temperature was found to be 106.7°K as compared to the published value 96.3°K, normalization being referred to 293°K. The Nb Debye temperature was found to be 56.5°K using the conductivities 6.14×10^6 mhos/meter (293°K) and 24.1×10^6 mhos/meter (77.3°K), normalized at 293°K.

The teflon dielectric was extruded upon the Nb center conductor. The Nb wire is not pure and contains 0.25 per cent impurities.[7] The 10-mil Nb wire was formed by sintering a metallic powder into an ingot and drawing to size. Slade and Smallman[8] have found that 3-mil samples of the same wire even after vacuum annealing are not exceptional superconductors. The maximum supercurrent that could be carried by the 3-mil wire before the Silsbee destruction of the supercurrent occurred was approximately 50 per cent of that which the Silsbee Hypothesis predicts.[9]

Future studies on the miniature line will be concerned with the Silsbee effect and the measurement of the conductivities from 293°K down to the superconduction critical temperatures. Also, the *X*-band dielectric properties of teflon at 4.2°K are to be investigated. The results of these studies in conjunction with a study of the high-frequency behavior of superconductors is expected to provide a basis for calculating the bandwidth of the line.

Conclusion

This investigation has demonstrated that a miniature superconducting line may be employed as a relatively lossless wideband coaxial line that is able to transmit nanosecond pulses with no apparent distortion. Furthermore, the experiments performed within this investigation in conjunction with the present-day knowledge of low temperature normal conductivity indicates the feasibility for designing a miniature low-temperature normal conducting line which would possess an attenuation several orders less than 293°K attenuation. One point that should be kept in mind is that if the losses vanish in a transmission line, the line geometry must be perfectly uniform if reflections within the line itself are to be avoided. Under lossless conditions, such internal reflections would not be damped by resistive loss, and the resulting multiple reflections could affect the signal-to-noise ratio of the transmission line; hence, under some circumstances in a given line, a combination of normal and superconductors would be appropriate, *e.g.*, inner conductor—super and outer conductor—normal.

Acknowledgment

The authors would like to acknowledge their discussions with R. L. Wigington and the ardent help of their graduate research assistant, A. J. Cummings. Also, the authors are indebted to their sponsor who supported their work within DA 49-170-sc-2618 and DA 18-119-sc-899.

[7] Per cents: Fe—0.02, Ni—0.01, W—0.01, Ta—0.05, Ti—0.02, Zr—0.02, O—0.05, N—0.05, C—0.02. Manufactured by Fanstell Metallurgical Corp., Chicago, Ill.

[8] A. E. Slade and C. R. Smallman, Arthur D. Little Co., Cambridge, Mass.; private communication with C. R. Smallman.

[9] D. Shoenberg, "Superconductivity," Cambridge University Press, Cambridge, Eng., 2nd ed.; 1952.

A 400-CPS Tuning Fork Filter*

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Summary—**The tuning fork filter as a new circuit component offers certain advantages in terms of frequency stability and large values of Q. An eleven-element comb filter is described in which each filter element has a bandwidth of 0.5 cps or a Q of 800, a 26-db insertion loss, and a 40-db rejection ratio. The frequency is constant to within 0.01 per cent, linearity to within one per cent, and Q to within ten per cent for a 40-db range of input. The temperature compensating effect of nickel used in the bimetallic fork is referred to an anomaly in its temperature-elasticity characteristic. The decoupling of the input-output electromagnetic transducers is described, and a novel method of Q adjustment in the form of shading coils is developed. The shape of the resonance curve shows that the tuning fork filter can be considered as equivalent to a single-tuned RLC circuit. The fork response at submultiples of its resonant frequency is due to slight harmonic generation in the drive circuit; the nonharmonic overtones are accurately predicted by cantilever theory. A 30-PPM position sensitivity is shown to be due to the change in the effective fork stiffness caused by gravity.**

I. Introduction

THE term "comb" is generally associated with a filter or filters used to cover a certain frequency band of interest with a predetermined resolution, and consisting of a number of equally and closely spaced filter elements. The ideal response of each element would be a single line on the frequency axis, and the frequency characteristic of an array of such elements would resemble the teeth of a comb. The usual approximation to this ideal frequency response is to allow each element of the comb to have a bandwidth equal to the frequency separation between elements, but this still allows a rather general class of transfer characteristics to be synthesized. When a comb filter is used for the detection of a signal in the presence of noise, its transfer characteristic can be adjusted to approximate the spectral density of the signal, and this is equivalent to cross-correlation.[1] A comb filter can also be used as a noise spectrum analyzer, or to detect the Doppler frequency shift of a signal. In the latter application, a tuning fork filter with a Q of 5000, described elsewhere,[2] would be a suitable component to provide the stability of the transmitted signal.

While comb filtering techniques have led to theoretical simplifications, their practical application in a given problem depends upon the availability of a stable, selective, and compact component for the filter element. These requirements make the use of conventional LC filters unattractive, and most comb filters use mechanical filter elements. Crystals and magnetostrictive devices have been used for comb filtering in the IF region, but they become impractical at the audio frequencies. Vibration galvanometers[3] and clamped blades or resonant reeds[4,5] have been used as filters from 27 to 1200 cps, but their Q's are between 50 and 400. The Q's of 1000 to 20,000 available from a tuning fork[6] make it the most practical component for a comb filter element in the audio-frequency region.

Historically, the mechanical resonance properties of the tuning fork have been appreciated in its acoustical applications since it was invented in 1711 by John Shore,[7] and this is still a current application.[8] As early as 1923, the accuracy and stability of this mechanical resonance led to an investigation of the tuning fork as the frequency selective component in an oscillator circuit,[9] and this use of the tuning fork has rivalled its acoustic applications.[10] A very recent and novel application of the tuning fork has been as a rate gyroscope.[11] While the use of a tuning fork as a filter has been suggested,[12] the paper by Wulfsberg[6] seems to be the sole description of the device in the literature. Of course, many of the references to tuning forks used in oscillator circuits are pertinent to their filter application.

II. Description of the Filter

The comb filter to be described uses 11 tuning forks to cover the frequency region from 397.5 to 402.5 cps. Each filter element, therefore, has a Q of approximately 800, or, more directly, a bandwidth of 0.5 cps. It is desirable that this bandwidth be stable to within 10 per cent, that is, 0.05 cps. In order to preserve this accuracy in the crossovers between adjacent filter elements, the tolerance for the resonant frequency is also 0.05 cps or

* Received by the IRE, May 31, 1960; revised manuscript received August 26, 1960.

† Hathaway Instruments, Inc., Denver, Colo.

[1] S. F. George and A. S. Zamanakos, "Comb filter ng for pulsed radar use," PROC. IRE, vol. 42, pp. 1159–1165; July, 1954.

[2] J. J. O'Connor, "An electrically-tunable tuning fork filter," *Electronics;* October 28, 1960.

[3] F. M. Gol'tsman, "L.F. vibration filters," *Zh. tekh. Fiz.*, vol. 24, pp. 1350–1353; July, 1954. (In Russian.)

[4] E. Fubini-Ghiron, "Electromechanical filters of variable frequency," *Alta Frequenza*, vol. 7, pp. 138–144; February, 1938. (In Italian.)

[5] G. Zelinger, "Tunable audio filters," *Electronics*, vol. 27, pp. 173–175; November, 1954.

[6] K. N. Wulfsberg, "A tuning fork filter," *J. Acoust. Soc. Am.*, vol. 22, pp. 847–849; November, 1950.

[7] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., p. 3; 1949.

[8] C. C. J. Addink, "Precise calibration of tuning forks," *Philips Tech. Rev.*, vol. 12, pp. 228–232; February, 1951.

[9] D. W. Dye, "The valve-maintained tuning fork as a precise time standard," *Proc. Roy. Soc. (London)*, vol. 103A, pp. 240–260; May, 1923.

[10] See for example, F. Haas, "A transistor-driven tuning fork frequency standard," *Toute la Radio*, vol. 23, pp. 282–283; September, 1956. (In French.)

[11] J. B. Chatterton, "Fundamentals of the vibratory rate gyro," *Trans. ASME*, vol. 78, sec. 1, pp. 123–125; January, 1956.

[12] Muirhead and Co., Ltd., "Miniature electrically maintained tuning-fork," *J. Sci. Instr.*, vol. 17, p. 134; May, 1940.

approximately 0.01 per cent or 100 PPM. The joint requirements of a relatively large Q and small frequency tolerance dictate the use of tuning forks for the filter elements. The 400-cps frequency of the comb is well within the practical range of, *e.g.*, 50 cps to 10 kc for tuning fork filters, but ironically, it proved difficult to reduce the basic tuning fork Q to the required value. The following sections describe the design of the various features and parameters of the tuning fork, many of which are peculiar to filter applications and not generally considered in oscillator circuits.

A. Transducers

While a tuning fork can be used as a two-terminal device in an oscillator circuit,[13] it would require a bridge circuit in a filter application.[14] As a four-terminal device with separate transducers for the drive and the pickup, there is a choice of electrostatic drive[15] with either electrostatic[16] or photoelectric[17] pickup, direct use of the piezoelectric effect with a quartz tuning fork,[18] or electromagnetic drive and pickup. The last method is used in this application because of its conventional design considerations, and the degree of coupling to the fork is easily adjusted. There are also certain disadvantages to this method which will be discussed below. Its use, of course, presumes a fork made of magnetic material.

The input signal is applied to the fork via wire-wound coils positioned near each fork tine. Since the electromagnetic attraction between the fork tines and the coils is independent of the polarity of the signal current, a magnetic bias is necessary in the drive circuit to avoid frequency doubling effects. A magnetic bias is also necessary in the pickup circuit so that the varying reluctance of the bias path, caused by the movement of the fork tines, will result in an EMF being generated in the pickup coil. These bias fields are supplied by small, U-shaped Alnico II magnets, and the arrangements of the several coils and magnets are shown in Fig. 1.

B. Rejection Ratio

While it is usual for a tuning fork used in an oscillator circuit to have one drive and one pickup coil, each with its own magnet and both mounted between the fork tines, this arrangement causes excessive off-resonance coupling which would lead to a reduced rejection ratio in a filter application. Therefore, the symmetrical arrangement of two outside drive coils and one inner pickup coil is used, as shown in Fig. 1. The winding orientation of the drive coils was selected to minimize the inductive coupling to the pickup coil, and it was verified that minimum coupling was obtained independent of the position of the drive coils, provided only that they were symmetrically placed on their respective magnets. The pole orientation of the drive magnets was then selected to produce an in-phase drive when the drive coils are excited in series. While the coupling was thus minimized, it was not reduced to zero. The residual inductive coupling causes an extraneous transmission which reaches a 10-kc maximum about equal to the fork transmission at resonance. This feed-through is cancelled by placing a 0.01 μf capacitor across the pickup coil. There is also capacitive coupling between the input and output coils because of the presence of the fork, and this is eliminated by grounding the fork and its base. This input-output coupling is one disadvantage of electromagnetic transducers. While rejection ratios of 50 to 60 db are available by 90° space orientation between the drive and pickup coils, 40 db is the practical maximum for the arrangement as shown in Fig. 1.

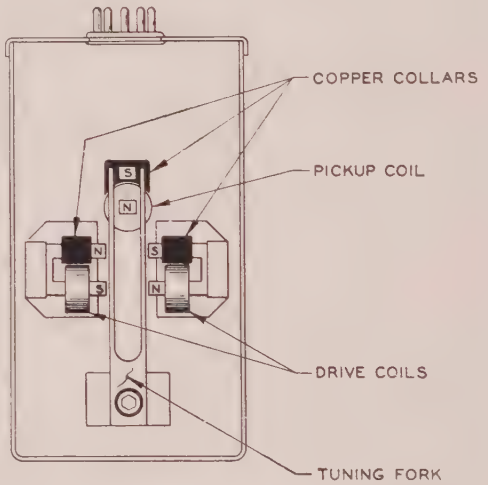

Fig. 1—Configuration of tuning fork filter.

C. Insertion Loss

Since the fork is isolated by drive and pickup amplifiers, its input and output impedances were not design parameters. Each coil has 13,000 turns of AWG 50 wire with a resistance of 12.5K. The input impedance at resonance is 28K, and the output impedance is 16K. Under this condition, the insertion loss is 26 db, that is, 20:1 voltage attenuation. Since the impedances are determined mainly by the coils, they can be designed for circuit considerations or minimum insertion loss. As an example, the insertion loss can be reduced to 12 db by using 600-ohm drive coils, or even less if lower input impedances are not objectionable. Reduced insertion

[13] E. G. Richardson and Y. L. Yousef, "An electrostatically maintained tuning fork," *Proc. U. Durham Phil. Soc.*, vol. 11, pp. 119–124; September, 1953.

[14] L. G. Bostwick, "Vibratory reed-controlled oscillator," U. S. Patent No. 2,583,542; January 29, 1952.

[15] P. Grivet, "An electrostatic method of maintaining the vibrations of tuning forks and bars," *Compt. rend. (Paris)*, vol. 213, pp. 231–233; August 11, 1941. (In French.)

[16] A. Bolle, "Comparison of the frequency of two tuning forks," *Ricerca Sci.*, ser. 2, vol. 2, pp. 697–702; December 15–31, 1938. (In Italian.)

[17] Cambridge Instrument Co., Ltd., "A tuning fork maintained by a photocell," *J. Sci. Instr.*, vol. 17, p. 72; March, 1940.

[18] L. D. Bryzzhev and V. N. Titov, "A simple clock with a quartz tuning fork," *Zh. tekh. Fiz.*, vol. 24, pp. 879–883; May, 1954. (In Russian.)

loss can also be obtained by reduced air gaps on the coil magnets[6] at the expense of frequency stability as described below.

D. Temperature Compensation

While the sharpness of the mechanical resonance of the tuning fork gives large values of Q, the resonant frequency is a function of temperature in the following manner. Using the equation for a cantilever,[19] which has experimentally been found to hold for a tuning fork, the fundamental frequency is:

$$f_1 = 0.55966(E/d)^{1/2}k/L^2, \quad (1)$$

where

f_1 = the fundamental frequency in cps,
k = the radius of gyration of the tine cross section in cm,
L = the length of the tine in cm,
E = Young's modulus in dynes/cm^2, and
d = density in grams/cm^3.

For a tine with rectangular cross section of width T in the plane of vibration, the radius of gyration is

$$k = T/(12)^{1/2}, \quad (2)$$

and the fundamental frequency is

$$f_1 = 0.16156(E/d)^{1/2}T/L^2. \quad (3)$$

Assuming that Young's modulus is temperature dependent, and that the linear dimensions will each follow the same temperature function, we may write

$$E = E(t) \quad (4)$$

$$T = K_1 y(t) \quad (5)$$

$$L = K_2 y(t) \quad \text{and} \quad (6)$$

$$d = \text{mass/volume} = M/K_3 y^3(t), \quad (7)$$

where all the K's are constants, t is the temperature variable, and $y(t)$ is the assumed temperature-expansion function. Substitution of (4) to (7) into (3) yields, after suitable manipulation,

$$f_1 = K_4[E(t)y(t)/M]^{1/2}. \quad (8)$$

Eq. (8) shows that a positive expansion coefficient would lead to a positive frequency coefficient. However, the frequency of a steel tuning fork decreases with temperature at a rate of approximately 100 PPM/°C because it is controlled by the decrease of Young's modulus. Eq. (8) also shows that the low expansion coefficients ($y(t) = K$) found in Invar-type alloys would not lead to low frequency coefficients because the stiffness coefficient would still govern.[20] However, advantage can be made of an anomaly in the temperature-elasticity function of nickel,[21] as shown in Fig. 2, in that the modulus increases over a certain temperature range when the nickel is less than 50 per cent magnetized. Various alloys such as Magnivar, Elinvar, and Ni-Span-C use nickel for balanced stiffness and expansion coefficients and frequency coefficients as low as ±5 PPM/°C have been obtained.[22]

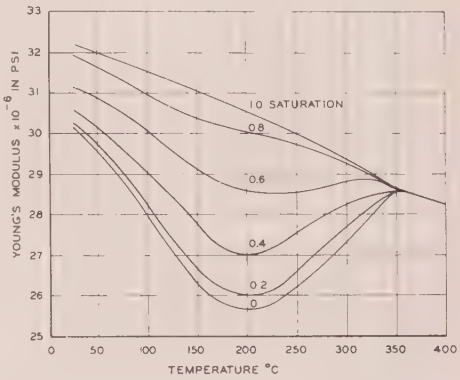

Fig. 2--The temperature dependence of Young's modulus as a function of magnetization in nickel. (From Siegel and Quimby.[21])

The use of a bimetallic fork leads to a slightly larger range of temperature compensation than that obtained with one of the monometals. A suitable nickel-steel alloy is chosen in which the minima of the temperature-elasticity curves are below −50°C, and a positive slope is thus obtained over the temperature range for tuning fork operation. This nickel-steel is then joined to a lamination of carbon-steel which has the usual negative temperature-elasticity slope. The ratio of the thicknesses of the two metals can be adjusted so that their combined moduli and expansion coefficients tend to cancel. With care, frequency coefficients of less than one PPM/°C can be obtained, and the total frequency change from −50 to +100°C can be held to ±50 PPM. The ultimate accuracy is limited by curvature of the resultant temperature function, and greater accuracy requires temperature controlled ovens.

In general, a low temperature coefficient can not be specified for a fork apart from its specific amplifier and drive system.[23] One reason for this is that below the Curie point the temperature coefficient of Young's modulus is a function of magnetization, as shown in Fig. 2, and the degree of magnetization of the fork depends upon the strength and spacing of the drive and pickup magnets. Further, the strength of the magnets, and consequently the "magnetic stiffness" of the fork as well as coupling of the transducers, are all functions

[19] P. M. Morse, "Vibrations and Sound," McGraw-Hill Book Co., Inc., New York, N. Y., p. 158; 1948.
[20] H. L. Clark and H. Johnson, "Standard frequency source," *Gen. Elec. Rev.*, vol. 47, pp. 42–46; May, 1940.
[21] S. Siegel and S. L. Quimby, "The variation of Young's modulus with magnetization and temperature in nickel," *Phys. Rev.*, vol. 49, pp. 663–670; May, 1936.
[22] H. W. Lamson, "A simple and precise standard of musical pitch," *J. Acoust. Soc. Am.*, vol. 7, pp. 51–55; July, 1935.
[23] E. Norrman, "Tuning fork stabilization," *Electronics*, vol. 13, pp. 15–17; January, 1940.

of temperature. The physical dimensions and spacings of the mechanical system, and such electrical parameters as the resistance of the drive and pickup coils, have some variation with temperature which enters into the fork's temperature coefficient of frequency. The main advantage of the bimetallic fork over the various monometallic forks is the fact that the compensation process can account for the resultant temperature variation from all causes.

E. Bandwidth or Q

The basic Q of the bimetallic 400-cps tuning forks was 1800 while the parameters of the comb filter required a Q of approximately 800. It would have been possible to obtain the required value of Q by reducing the drive magnet spacing from 0.040 inch to 0.010 inch. This reduction of Q is caused by increased eddy current losses in the tines, which in turn are caused by the increased magnetic field around these moving conductors. While this scheme would have reduced the insertion loss, it would not provide sufficient isolation between the fork and the transducers. The result would have increased the nonlinearities between the resonant frequency and drive amplitude as described below.

A better method of Q reduction is to withdraw energy electrically from the vibrating system, the amount of this energy being proportional to the amplitude of vibration. This can be done by a coil and magnet arrangement similar to that of the pickup transducer. (The pickup transducer itself can not be used because excessive attenuation of the output signal or increased insertion loss will result when the coil is loaded due to its internal impedance.) The vibrating tines will cause an EMF to be generated in the coil due to the varying reluctance of the magnetic circuit, and the amount of energy withdrawn from the fork by this EMF can be controlled by loading the coil with the potentiometer. This has proved to be a convenient method of Q adjustment in other applications where a 20 per cent variation of Q has been obtained. In the present application, however, the necessary Q reduction required the most efficient use of the fixed volume for the coil, and this meant a coil of unity space factor, that is, a solid copper. The effectiveness of this single turn coil is determined only by the magnitude of ac flux which links it, and since it is in the form of a solid copper collar, it was shaped to clear the coils on the pickup and drive magnets, as shown in Fig. 1. By using three such coils, the necessary Q reduction was obtained. It should be noted that these copper collars did not increase the insertion loss or add phase shift between the drive and pickup circuits. Further, they did not affect the temperature compensation of the fork.

F. Resonant Frequency

Eq. (3) can be used to calculate the dimensions of a tuning fork for a given resonant frequency, but it is only accurate to a few per cent. Greater accuracy can be obtained by additive corrections to the tine thickness and the tine length,[24] but these amount to empirically derived values to account for the fork geometry. Further, the effective stiffness added to the system by the attraction of the magnets makes it difficult to predict the resonant frequency to the required accuracy of 0.01 per cent. Therefore, the final frequency of each fork is empirically adjusted either by adding mass, in the form of solder to the end of the tines, or by reducing the mass by removing metal from the ends of the tines. The range of adjustment is such that all the frequencies of the comb (397.5 to 402.5 cps) can be obtained from forks of the same original size.

Besides the inaccuracies of predicting the resonant frequency, there is also a matter of balance. Even though a fork is made by careful machining processes and is aligned on its base to close physical tolerances, the tines will generally resonate at two slightly different frequencies caused by electrical, mechanical, or magnetic dissymmetries. This feature was investigated by increasing the unbalance in an attempt to derive tuning fork filter characteristics equivalent to those of double-tuned or stagger-tuned circuits. It was possible to separate the resonances of the tines and to derive two peaks in the transmission characteristic. However, under these conditions the fork and its base formed one vibrating system, and the frequencies of vibration depended upon how the base was mounted on the bench. This unstable operation of the fork is thus undesirable, and the vibration fed to the base should be minimized in order to make the fork frequency independent of its mounting. This is accomplished in the balancing process which is actually part of the final tuning described in the previous paragraph. It should be noted that the balanced condition is obtained by a combination of mechanical and magnetic parameters.

III. Transfer Characteristics

While the filter application of the tuning fork is a logical extension of its use in oscillator circuits, certain filter parameters become important. The exact shape of the transmission characteristic both near to and removed from resonance determines the transient response, averaging effect, and noise reduction of the filter. Input-output linearity is a general requirement as well as independence of the resonant frequency from drive amplitude. These and other items will be investigated in some detail.

A. Resonance Curves

By considering the tuning fork as a simple, second order system with mass, stiffness, and friction, its translational motion can be described by a linear, second order differential equation. The solution of this equation leads to the simple resonance curve in the fre-

[24] A. Winkelmann, "Handbook of Physics," Springer, Berlin, vol. 2, p. 348; 1909. (In German.)

quency domain. A more complete analysis of its vibration, which accounts for the fork's geometry, proves to be untractable, but it does raise a question concerning the fork's frequency response. To investigate this, the experimentally measured frequency response of a tuning fork filter can be compared to the theoretical resonance curve calculated in the following way. The frequency response of a simple, second order system can be written as

$$F(\omega) = 1/(1 + j2Q_0\delta) \qquad Q_0 \gg 1,\ \delta \ll 1 \tag{9}$$

where

$$\delta = (\omega - \omega_0)/\omega_0, \tag{10}$$

and Q_0 is the ratio of the energy stored per cycle to the energy dissipated per cycle at the resonant frequency ω_0. Assuming that the usual relationships hold between these parameters, we may write

$$Q_0 = \omega_0/\Delta\omega, \tag{11}$$

where $\Delta\omega$ is the 3-db bandwidth. Therefore,

$$Q_0\delta = (\omega - \omega_0)/\Delta\omega = x, \tag{12}$$

where x is the frequency variable relative to the resonant frequency and normalized to the bandwidth. Eq. (9) can then be written as

$$F(\omega) = 1/(1 + i2x), \tag{13}$$

$$|F(\omega)|^2 = 1/(1 + 4x^2). \tag{13a}$$

Eq. (13a) is plotted on the graphs of Fig. 3 with a resonant frequency of 400 cps and a bandwidth of 0.5 cps so that the curve may be compared to the experimental data points collected from the central filter of the comb. Fig. 3(a) is an enlarged section of the graph to show how well the data points fit the theoretical curve near resonance, and Fig. 3(b) shows that this agreement is also found on the skirts. While some asymmetry might be expected from considerations discussed below, there is no evidence of it in the data of Fig. 3. Therefore, the simple resonance curve of (9) or (13) can be used to describe the tuning fork filter characteristic, and the most useful definition of Q is given by (11).

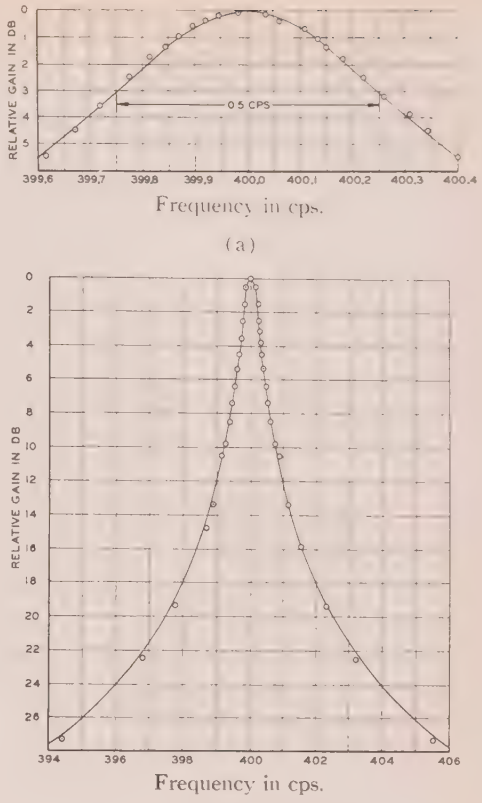

Fig. 3—Resonance curve of tuning fork filter.

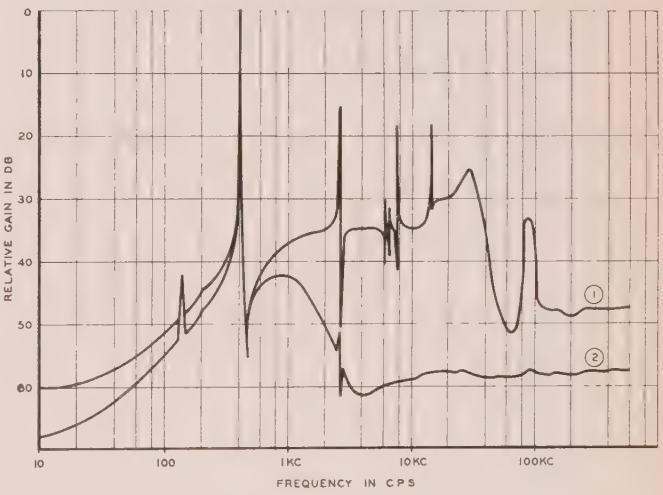

Fig. 4—Transfer characteristic of tuning fork filter.

B. Frequency Response

The frequency response of the tuning fork filter from 10 cps to 600 kc is shown on Fig. 4 normalized to the 26-db insertion loss. Curve 2 is the response with the 0.01-μf capacitor across the pickup coil which gives the 40-db rejection ratio discussed above. Curve 1 is the response without this capacitor, and it contains several interesting features which will be discussed in turn.

The most interesting feature of the response curve is the sharp cusp near the 400-cps resonant peak. This cusp can be explained as follows: assume that there are several parallel paths between the input and output terminals of the filter, and that the output signal is the addition of the several parallel signals. One signal is transmitted by the fork and follows its resonance characteristic. Another signal is the feed-through caused by the inductive and/or capacitive coupling between the drive and pickup coils. The fork signal is capacitive at frequencies slightly higher than the resonant frequency, while the increasing nature of the feed-through signal in this region indicates that it is inductive. The algebraic addition of these signals results in smaller amplitudes over this frequency region caused by the relative phase angles of the signals. Since the fork sig-

nal is rapidly decreasing in amplitude and increasing in phase lag, there is one frequency at which it is most effective in cancelling the feed-through signal. This minimum resultant amplitude of the two signals in the cusp and its sharpness is caused by the fork's selectivity. It should be noted that this cusp can appear on the low-frequency side of resonance by a sign change of the feed-through signal relative to the fork signal.

Another point of interest is the small peak at 133 cps. This is not truly part of the transfer characteristic and represents a nonlinearity in that the peak is caused by a 400-cps output signal. There is a third harmonic of the 133 cps drive signal generated in the magnetic circuit, and although it is small, it is capable of exciting the fork into its fundamental mode of vibration. There is a hint of this phenomenon in the curve at 200 cps, which means a small amount of second harmonic is also generated in the drive circuit.

The resonant peak and cusp near 2.5 kc, which appears on both curves, is caused by the drive signal exciting the fork in its first overtone. Similar peaks near 7 and 13 kc indicate the second and third overtones, respectively. These peaks were identified as overtones by locating the expected node pattern along the tines with a stylus. The ratios of these overtones to the fundamental are 6.4, 18.1, and 34.8. The theoretical overtones of a cantilever[19] are 6.27, 17.55 and 34.39. This is very close agreement, and the fact that the fork frequencies are a few per cent higher than the cantilever frequencies can be attributed to the extra tine thickness near the crotch. Further, the fork frequencies are somewhat determined by the stiffness added to the system by the several magnets.

This agreement between tuning fork overtones and cantilever theory presents new evidence for an old controversy[25] of whether a tuning fork should be considered as a set of clamped beams (cantilevers) or a U-shaped bar. The latter approach has intuitive appeal in its development of the tuning fork from a simply supported free-free bar; the nodes which are approximately one-quarter of the way in from the ends of the bar move towards the center or stem as the bar is bent into the form of a U. This theory hypothesizes the existence of two nodes in the stem of the fork, and this has been verified experimentally.[26] The presence of these nodes indicates that there is vibration of the stem or crotch of the fork which is complementary to the tine vibration, and it is this vibration which accounts for the audio amplification when a vibrating tuning fork is placed on a resonance box. This base vibration also indicates that care must be taken in the mounting of a tuning fork in precision instruments in order to isolate the fork's vibration. However, it is cantilever theory that gives the quantitive prediction of the tuning fork's fundamental frequency in terms of its geometry, as shown by Eq. (1), and it further predicts the ratio of the overtones to this fundamental frequency.

Fig. 4 also shows a double peak near 6 kc which is not a fork overtone as established by stylus probing. The investigation of this double peak did not identify its cause, and it is probably a resonance of some portion of the mechanical system comprising the fork, its base, the magnets, etc. The broad peak near 30 kc is caused by self-resonance of the coils. The isolated peak near 80 kc was not investigated as it did not appear to be associated with the fork and may have been caused by the signal generator or other extraneous sources. All of these features above the first overtone are not present in the filter transfer characteristic, as shown in the lower curve 2 of Fig. 4, because of the capacitor across the pickup coil.

C. Nonlinearities

Perhaps the greatest difference between oscillator and filter applications for a tuning fork is that, in the former, various nonlinearities can be avoided by having a constant amplitude drive. A tuning fork filter, on the other hand, should have a dynamic range of operation over which the filtering action is essentially linear. Fig. 5 shows the drive voltage on a logarithmic scale, in order to investigate the several fork parameters over a dynamic range of the input, as the voltage ratio. Again, the data were taken from the central comb filter element which has a resonant frequency of 400.000 cps and Q of 800.0 for a one-volt drive. The upper curve of Fig. 5 shows that the resonant frequency decreases slightly, about 35 PPM, at the smaller drive amplitudes, and this decrease could be caused by slight cubic curvature of the stress-strain curve near its origin. However, the resonant frequency is substantially con-

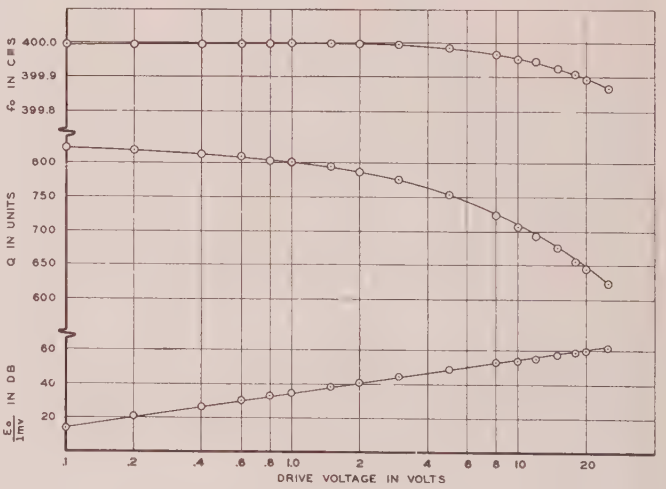

Fig. 5—Resonant-frequency, Q, and output voltage as functions of input voltage of a tuning fork filter.

[25] P. T. Young, "The vibrations of a tuning fork," *Science*, vol. 54, pp. 604–605; December 16, 1921.

[26] S. E. Michaels, "Tuning Fork," U. S. Patent No. 2,247,960; July 1, 1941.

stant over a 40-db range and deviates 100 PPM or more from its nominal value only at drive amplitudes greater than 10 volts.

The decrease of the resonant frequency at the larger drive amplitudes is expected and is caused by the restoring force being a function of tine position. Part of this nonlinearity is caused by curvature of the stress-strain curve and to the mechanical relationship between tine deflection and bending moment. To this extent it is inherent in the tuning fork. In an electrically driven tuning fork, however, the electromagnetic forces acting on the tines are a function of the instantaneous air gap dimension, that is, tine position; they thereby affect the resonant frequency by changing the effective stiffness of the fork. Rayleigh and others[27] predict such a decrease of resonant frequency with increased vibration amplitude from the differential equations of motion by assuming a polynomial expression for the restoring force as a function of tine position. The solution of these nonlinear differential equations requires the assumption that the fork's vibration be simple harmonic motion. The data collected for Fig. 5 verify this assumption in that the output signal was a pure sine wave even at the largest vibration amplitudes. A further point of interest is that Mallett[27] predicts that increasing drive amplitude will distort the filter characteristic, eventually making it double-valued at the low-frequency side of resonance. This has been observed by overdriving a fork. Over a certain band of frequencies there were two discrete amplitudes of vibration possible for a given drive amplitude, the choice being determined by the direction in which the frequency was approached.

The middle curve of Fig. 5 shows that the Q continually decreases with increasing amplitude but is within a 10 per cent tolerance over the 40-db range of drive voltages less than 10 volts. This decrease of Q with increasing amplitude is caused, in part, by the increased internal friction, but is also caused by the increased hysteresis and eddy current losses caused by the magnetic transducers. Both the decrease of resonant frequency and Q with increasing drive amplitude are aggravated by magnetic effects and, therefore, the input and output transducers should be loosely coupled to the fork. It is for this reason that the necessary reduction of Q was not obtained by smaller air gap spacings.

On the lower curve of Fig. 5, the output voltage at resonant frequency has been plotted as db relative to one millivolt, in order to indicate that it is proportional to a power of the input voltage. By a least square curve fit, this power has been calculated to be 0.97 with a 25.5-db insertion loss at one volt drive. By setting the amplifier gain at 26 db, the input is exactly equal to the output at 7.5 volts drive, and the largest error over the 40-db range is less than one per cent at 10 volts drive. This increase of insertion loss with increased drive is related to the decrease of Q, and it does not cause any distortion in the output waveform which is a pure sine wave.

D. Position Sensitivity

The data of Fig. 6 show that the resonant frequency of a tuning fork depends upon its orientation relative to the gravity vector, which is shown as a reference in each of the sketches. In order to obtain the data of Fig. 6 with the necessary accuracy, the tuning fork was put into self-oscillation, and the oscillation frequency was measured as a function of fork position. The sinusoidal curves were calculated from the data by a least square curve fit.

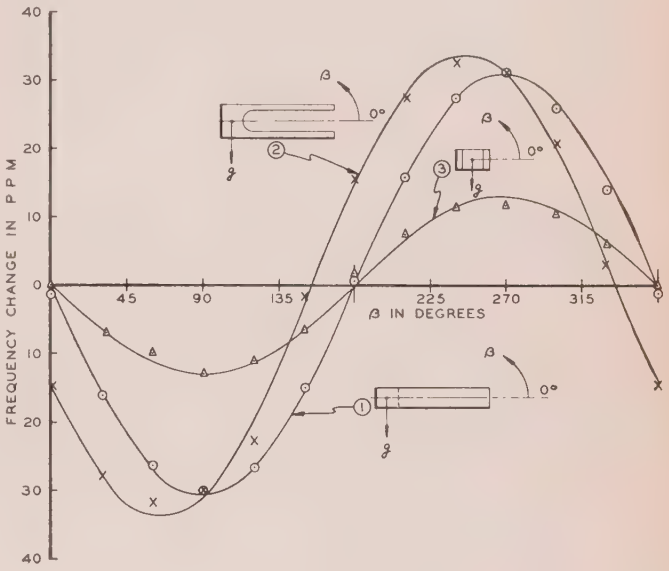

Fig. 6—Position sensitivity of a tuning fork.

For curve 1, the fork was rotated about its heel with its normal to the plane of tine vibration kept in a vertical plane. These data are as expected,[22] being sinusoidal with a minimum frequency when the tines are vertically up and a maximum frequency when the tines are vertically down. The total frequency variation is seen to be ± 30 PPM. For curve 2, the fork was rotated about its heel maintaining its plane of vibration vertical. This function is also seen to be sinusoidal, but its peak values of 34 PPM occur at 25° off the vertical. This was not expected, nor is it believed to be general. It is recalled that there are three magnets, two outside the tines and one between them as shown in Fig. 1. Any differential displacement of the tines or asymmetry in the magnetic fields which would change the effective stiffness of the fork could account for the extra frequency change of a few parts per million at the nonvertical position.

For curve 3, the fork was rotated about its longitudinal axis which was horizontal. Again a sinusoidal variation was observed, but none was expected.[23] Its magnitude, however, is only ± 12 PPM, and this could also

[27] E. Mallett, "Distortion of resonance curves of electrically-driven tuning forks," *Proc. Phys. Soc. (London)*, vol. 39, pp. 334–358; June, 1927.

be caused by the deflection of the limber tines in the several magnetic fields.

Curves 2 and 3 may represent anomalies peculiar to this particular fork configuration, but curve 1 represents a position sensitivity which is inherent in the tuning fork. In fact, the resonant frequency of a tuning fork is affected by any longitudinal acceleration vector, gravity being a special case. It has been verified that this positional sensitivity is independent of the fork vibration amplitude but is linearly proportional to the applied acceleration. These phenomena can be explained by the following simplified model. Each tuning fork tine is considered as a concentrated mass m which undergoes a radial vibration through an angle θ at a distance d from its node in the base. The effective spring constant is s, and all the damping terms are included in R. In the absence of applied acceleration, the differential equation of the system can be written by equating the sum of the torques around the node points to zero; that is,

$$md^2\ddot{\theta} + R\dot{\theta} + s\theta = 0. \tag{14}$$

In a high Q system (low damping) the natural frequency ω_0 is given as

$$\omega_0^2 = s/md^2. \tag{15}$$

When an acceleration a is applied parallel to the longitudinal axis of the fork, it will produce a force ma which acts through a distance arm $r \sin \theta$ to produce a torque T_a which is equal to

$$T_a = mad \sin \theta. \tag{16}$$

Since the vibration angle θ is small, even when the fork is overdriven,

$$\sin \theta \approx \theta \tag{17}$$

and

$$T_a \approx mad\theta. \tag{18}$$

When the fork tines are vertically up, the gravity vector will oppose the spring restoring force, and therefore this torque is added to (14) as

$$md^2\ddot{\theta} + R\dot{\theta} + s\theta - mda\theta = 0, \tag{19}$$

$$md^2\ddot{\theta} + R\dot{\theta} + (s - mda)\theta = 0. \tag{19a}$$

The resonant frequency ω_a is, therefore,

$$\omega_a^2 = (s - mda)/md^2 = \omega_0^2 - a/d \tag{20}$$

$$a/d = \omega_0^2 - \omega_a^2 = (\omega_0 - \omega_a)(\omega_0 + \omega_a). \tag{21}$$

The position sensitivity $(\omega_0 - \omega_a)$ is, therefore,

$$(\omega_0 - \omega_a) = a/d(\omega_0 + \omega_a). \tag{22}$$

This derivation has indicated the basic mechanism of position sensitivity; the acceleration vector changes the effective spring constant of the system which modifies the resonant frequency. Only the longitudinal component of the acceleration vector is effective in producing this change, and this explains the sinusoidal nature of the curves of Fig. 6. As a modification of the spring constant, the frequency change is independent of the fork vibration amplitude as shown in (19a). However, the frequency change is linearly proportional to the magnitude of the acceleration vector as shown in (22). Eq. (22) also indicates that the frequency change is less for higher frequency forks, and this has been verified with 60-cps forks having a frequency change of a few hundred parts per million, while that of 6000-cps forks is less than one part per million. However, (22) indicates that this relationship is proportional to the first power of the fork frequency $(\omega_a \approx \omega_0)$; limited investigation of other forks indicates that a power between 1.5 and 2 better describes the experimental data. The reason for discrepancy is that the fork frequency and its position sensitivity are complex functions of the fork geometry which had not been adequately included in the simplified model used to derive Eq. (22).

While this position or acceleration sensitivity may seem like a disadvantage of the tuning fork, its magnitude is actually very small and is measured in parts per million. The effect of a one-dimensional acceleration of large amplitude and long duration can be minimized by orienting the fork's longitudinal axis perpendicular to this acceleration. This sensitivity can be eliminated from a tuning fork frequency standard by driving two forks mounted back to back on the same longitudinal axis with a common amplifier.[28] On the other hand, this position sensitivity would allow the tuning fork to be used as a reference vertical or accelerometer by comparing the frequencies of such back-to-back forks which are maintained by separate amplifiers.[29]

IV. Conclusions

Of the several points of theoretical interest, perhaps the most significant one is the accurate prediction of tuning fork overtones from cantilever theory. This is further emphasized by the close agreement between the fundamental frequency of the tuning fork and the formula based on cantilever geometry. However, this does not completely resolve the controversy as there is considerable intuitive appeal in the development of the tuning fork from a U-shaped bar. Also, this approach leads to the important considerations of node points and base vibration. The experimental observation of a sinusoidal output, even when the fork is heavily over-driven, validates this necessary assumption for the solution of the nonlinear differential equations of fork vibration.

[28] W. J. Holt, "Frequency reference for guidance system," *Electrical Design News*, vol. 5, pp. 38–39; February, 1960.

[29] B. J. Blasingame, "Accelerometer," U. S. Patent No. 2,928,668; March 15, 1960.

This allows the analytical description of certain fork characteristics if the empirical function between the restoring force and tine position is measured. Further corroboration of these equations is the observation of two vibration amplitudes over a frequency band for an over-driven fork.

Of more practical significance is the agreement between the measured filter transfer characteristic and the universal resonance curve. The equivalent circuit for a tuning fork filter can, therefore, be taken as a simple resonant circuit, which has the same resonant frequency and Q as the tuning fork. If the phase function of the filter transfer characteristic is important, the extraneous coupling between the input and output transducers will have to be minimized in order to eliminate the cusp. This can be done with space oriented coils or nonelectromagnetic transducers. In either case, these transducers should be loosely coupled to the fork, even at the expense of increased insertion loss, in order to reduce the variation of resonant frequency, Q, and output nonlinearities at the larger drive amplitudes.

The various fork defects and nonlinearities were presented for the purpose of completeness, but it should be noted that their magnitudes are small. The audio-frequency filter with a stability of 100 PPM and a Q of 800 is essentially linear over a 40-db range of input.

V. Acknowledgment

The author wishes to express his appreciation to D. O. Ballinger, G. V. O. Cook and W. Plywaski for their work in the early stages of the development of tuning fork filters.

The Cyclotron Resonance Backward-Wave Oscillator*

K. K. CHOW†, student member, ire and R. H. PANTELL†, associate member, ire

Summary—**An electron injected into a longitudinal magnetic field with some initial transverse motion will rotate at the cyclotron frequency in the transverse plane. If an RF electric field, which is polarized in the transverse plane and which oscillates at the cyclotron frequency, is present, a cumulative energy interchange between the electron and the field will occur. This principle of cyclotron resonance interaction has been applied to the construction of a backward-wave oscillator and amplifier which requires no slow-wave circuit.**

The elimination of the slow-wave circuit provides several advantages, particularly if one is interested in the generation of energy at millimeter wavelengths. At these short wavelengths, it becomes extremely difficult to fabricate a slow-wave circuit, and, in addition, the interaction is weak because the fields decay exponentially away from the conducting surface. The present tube has approximately 100 times the transverse area of a helix operating at a comparable beam voltage and has strongest RF fields in the region of the electron beam.

The tube has been constructed for operation at S band, and the circuit consists of sections of standard S-band rectangular waveguide with a field pattern of the TE_{10} mode. An unbunched beam with some initial transverse motion is injected into a constant, longitudinal magnetic field. Bunching occurs in the interaction region, and, for sufficiently high beam current, oscillation power is produced at the gun end of the tube. The frequency of oscillation is determined by the magnetic field strength; therefore, the device can be tuned by changing the current in a solenoid. Tuning is continuous over a range from 2.1 kmc to 4 kmc. Highest peak power output is over 15 watts at 6 kv beam voltage and the highest efficiency obtained is 1.7 per cent at 3.5 kv beam voltage. Starting current varies from 4 to 40 ma over the frequency range, whereas the beam pitch angle remains roughly constant at 9°. Harmonic contents up to the fourth have been observed with the second harmonic power about 60 db below the fundamental. Backward-wave gain varies from 20 to 2 db over the frequency range, and amplification occurs over a band of about 7 mc centered at the cyclotron frequency.

Introduction

RECENTLY, a number of investigators have shown an interest in the possibility of obtaining interaction between a low-energy electron beam and RF electromagnetic fields propagating at a phase velocity at, or greater than, the free space velocity of light.[1–3] This has been termed "fast-wave interaction," as distinguished from the conventional traveling-wave tube interaction which involves waves propagating slower than the velocity of light. The reasons for showing such an interest are:

1) The tube can be very much simplified in that the circuit can be an unloaded waveguide or a coaxial or a multiwire transmission line. The difficulty of fabricating a complicated structure whose periodicity is less than the free-space wavelength is therefore eliminated. This is particularly advantageous at short wavelengths.

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† Microwave Lab., W. W. Hansen Labs. of Physics, Stanford University, Stanford, Calif.

[1] H. Kleinwachter, "Excitation of electromagnetic fields by current waves," *Arch. Elektrotech.*, vol. 6, pp. 376–378; September, 1952.

[2] R. Müller, "Space harmonics in electron beams," *Arch. Elektrotech.*, vol. 10, pp. 505–511; December, 1956.

[3] E. J. Gorn, "Travelling-wave electron reaction device," U. S. Patent No. 2,591,350; April, 1952.

2) Since the interaction circuit is unloaded, coupling and matching problems are reduced, for there is no required transition from the interaction circuit and the normal transmission circuit.
3) Interaction occurs in the region of the strongest RF field, as opposed to the slow-wave circuit case, in which the strongest fields are at the conducting surface. Again, this is an important consideration for the generation of millimeter waves.

The particular scheme that has been employed is to inject an electron beam from a field-free region into a standard rectangular waveguide immersed in uniform longitudinal dc magnetic field. The electrons rotate at the cyclotron frequency and interact with the dominant waveguide mode (the TE_{10} mode). Provided that appropriate synchronism and bunching conditions are satisfied, a cumulative interaction will occur and oscillation or amplification will be obtained. The advantages of this particular scheme over other "fast-wave" approaches have been discussed in some detail by Pantell,[4] and will not be repeated here.

Synchronism and Bunching

For the interaction to be cumulative, it is necessary than an electron should see the same phase of RF field all along its trajectory. Further, bunching must occur so that more electrons are decelerated than accelerated. Consider Fig. 1(a). Near the guide axis a circularly polarized component of the field propagates as

$$e^{j\omega t - j(\omega/v_p)z - jp\phi},$$

where

ω = angular frequency of the RF fields,
v_p = phase velocity of the RF fields,
z = longitudinal direction in which the fields propagate,
p = number of variations of the RF field pattern in the azimuthal direction, and
ϕ = azimuthal coordinate.

For an electron to see the same phase of the RF fields, it is necessary that

$$\omega t - \frac{\omega}{v_p} z - p\phi = \text{constant},$$
$$\text{for } z = \pm \dot{z}_0 t \quad \text{and} \quad \phi = \dot{\phi}_0 t, \tag{1}$$

where

$\dot{z}$ = dc longitudinal velocity,
$\dot{\phi}$ = dc angular frequency = cyclotron frequency = ηB,
η = magnitude of charge-to-mass ratio of an electron,
B = dc longitudinal magnetic field.

The $\pm$ sign indicates that RF fields can be propagated either forward or backward.

[4] R. H. Pantell, "Electron Beam Interaction with Fast-Waves," *Proc. Symp. on Millimeter Waves*, Polytechnic Inst. of Brooklyn, N. Y., vol. 9, pp. 301–312; April, 1959.

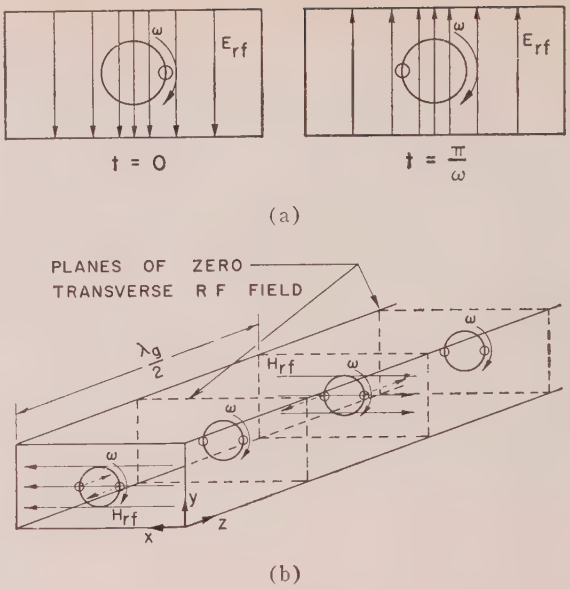

Fig. 1—(a) Synchronism condition. At a certain instant, $t=0$, an electron rotating in the transverse plane at an angular frequency ω sees a decelerating E_{RF}. Half a cycle later, $t=\pi/\omega$, E_{RF} reverses polarity and the electron rotates through half of its circular path so as to continue to deliver energy. (b) Bunching of electrons. Dotted arrows show directions of bunching force due to the cross product of H_{RF} and transverse electron velocity. Electrons which are 180° apart in the transverse plane will bunch $\frac{1}{2}\lambda_g$ apart in the Z direction.

Solving (1), we have

$$\omega = \frac{p\dot{\phi}_0}{1 \pm (\dot{z}_0/v_p)} = \frac{p\eta B}{1 \pm (\dot{z}_0/v_p)}. \tag{2}$$

For dominant waveguide modes, $p=1$, and with $\dot{z}_0 \ll v_p$,

$$\omega \doteq \eta B. \tag{3}$$

Thus, (3) establishes the synchronism condition which is also the tuning characteristic of the tube.

Before bunching occurs, electrons are distributed uniformly along the surface of a cylinder rotating at the cyclotron frequency. Under such a condition, there will be no net energy interchange between the electron beam and the electromagnetic fields, for there are as many electrons in an accelerating field as there are in a retarding field. Therefore, it is necessary to accomplish RF bunching. In a strong dc longitudinal magnetic field, electron motion in the transverse plane is inhibited by the presence of this field. It is only along the longitudinal direction that the electrons can move with relative ease. For a TE wave, the only component of RF field producing a longitudinally directed force is H_{RF} in the transverse plane. As shown in Fig. 1(b), H_{RF} interacts with the dc azimuthal electron velocity to provide longitudinal bunching. Electrons which are 180° apart in the azimuthal position in a transverse plane are moved in opposite directions. Eventually electrons which are separated by 180° are bunched $\lambda_g/2$ apart. After bunching has occurred, the electrons are in the form of a helix with a pitch equal to the guide wavelength and rotating at the cyclotron frequency. It is then possible to have

the majority of the electrons in a position and moving in a direction so as to deliver net energy to the circuit. Energy interchange occurs between the transverse motion of the electrons and the transverse RF electric field.

It is interesting to note that in the traveling-wave tubes, it is the E_{RF} that does the bunching and the extraction of energy. However, in the case of cyclotron resonance, bunching is provided by H_{RF} and energy extraction is provided by E_{RF}.

The tube will support amplification in both the forward and backward directions. Of course, if tube current is high enough, backward-wave oscillation will occur, since the feedback mechanism is inherently built-in. The use of this device as a forward-wave amplifier is limited because it is extremely narrow-band.

Further details regarding the theory of interaction and tube performance are now being prepared as a small-signal analysis.

Description and General Behavior of the Tube

Fig. 2 is a photograph of the tube, together with the solenoid that produces the longitudinal magnetic field. In operation, this solenoid is over the interaction section as shown in Fig. 3(a), which is a diagrammatic view of the essential features of the tube. The axial magnetic field distribution is illustrated in Fig. 3(b).

The gun produces a hollow beam which passes through a radial magnetic field at the anode gap and then through a "neck" into the main body of the tube. The outer shell of the gun and the anode structures are made of magnetic materials. A small auxiliary coil is placed on the central anode rod to produce the dc radial magnetical field at the anode gap. The beam is thereby given an angular velocity upon passing through the anode gap, so that most of the electron energy is converted to transverse kinetic energy; the ratio of transverse to longitudinal energy can be adjusted by controlling the amount of current into the auxiliary coil. It is desirable to have as much of the electron energy in the transverse motion as possible because:

1) Energy delivered to a circuit is extracted from the kinetic energy because of transverse motion.
2) The bunching force is proportional to the transverse velocity.
3) For a given physical length of interaction region, the interaction length along a beam trajectory is inversely proportional to the dc longitudinal velocity, which is small for a given beam voltage, if the transverse energy is large.

After passing the anode gap, the electrons do not rotate about a common axis with respect to the center line of the cathode. Rather, each electron rotates about its own axis, but since the RF field is essentially uniform across the beam area, it is not necessary that electrons have a common axis of rotation. The "neck" is a waveguide below cutoff for the tube and has a neck coil wound on it for beam focusing. However, it had been found that energization of this coil in either direction reduced the oscillation power and hence the neck coil was not employed in operation. The beam then enters the main magnetic field and picks up synchronism frequency as it reaches the plateau of the main field, and interaction with a waveguide mode occurs as described above. The plateau is over $8\frac{1}{2}$ inches long (less than one per cent variation) and is the result of the main solenoid and series-connected compensation coils. The output power is critically influenced by the positioning of these compensating coils; oscillation could be entirely eliminated by relatively minor changes in their location.

The tube consists of three sections of standard rectangular S-band waveguide: the output waveguide sec-

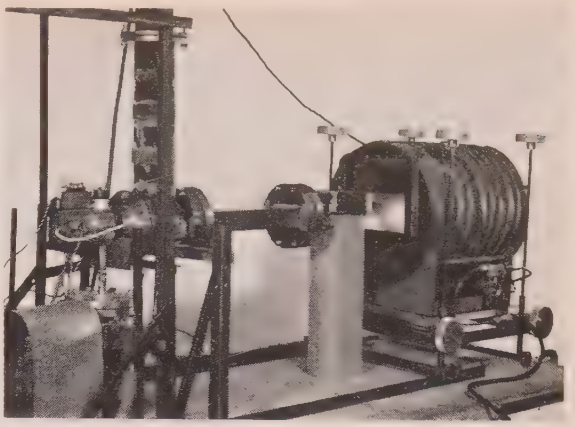

Fig. 2—Photograph of the cyclotron resonance fast-wave tube.

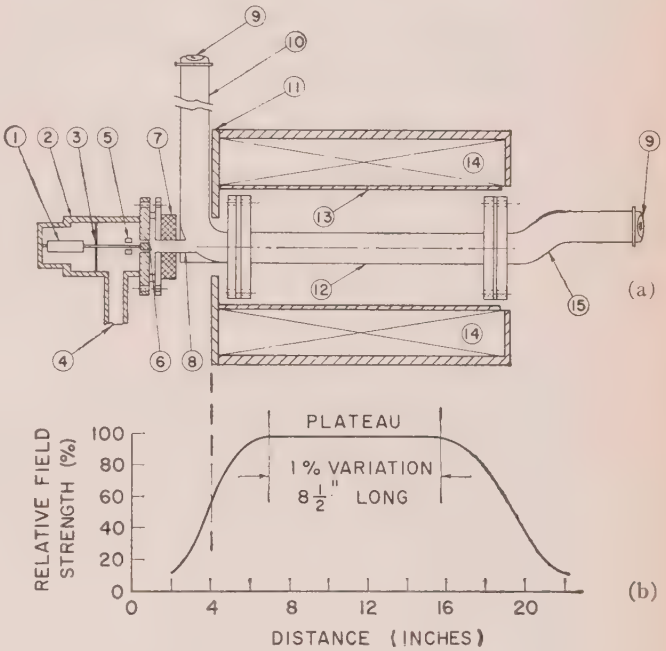

Fig. 3—(a) Cross-sectional view of the essential parts of the tube. (1) Auxiliary coil, (2) Gun envelope, (3) Vacuum seal, (4) Pump-out, (5) Cathode, (6) Anode gap with radial magnetic field for turning the beam, (7) Neck coil, (8) Neck, (9) Window, (10) Waveguide output, (11) Magnetic shield, (12) Interaction section, (13) Compensating coils, (14) Main solenoid, (15) Collector.

tion, the interaction section, and the collector section. The collector section is curved so as to avoid a direct optical path between the gun and the glass vacuum window. With the main solenoid on, no current was collected by the interaction section, *i.e.*, all the electrons entering the uniform magnetic field region reached the collector. The tube was pulse-operated at 60 pulses per second with a pulse length of about 10 μsec.

Experimental Results

A. Tuning Curve

Fig. 4 gives the tuning characteristic of the tube in the frequency range from 2.1 to 4.0 kmc. The theoretical curve is a plot of cyclotron frequency vs solenoid current. Cyclotron frequency was calculated by measuring flux density with a Hall effect probe that had been calibrated at 2.6 kmc. The discrepancy between the two curves of Fig. 4 can be attributed to the limited accuracy of the probe and calibration of the gaussmeter. The experimental curve consists of points taken at two beam voltages, 1.5 and 3 kv, and there is no observable difference between these two sets of data. The lower limit of frequency is set by cutoff of the waveguide, and the upper limit is set by power dissipation in the solenoid.

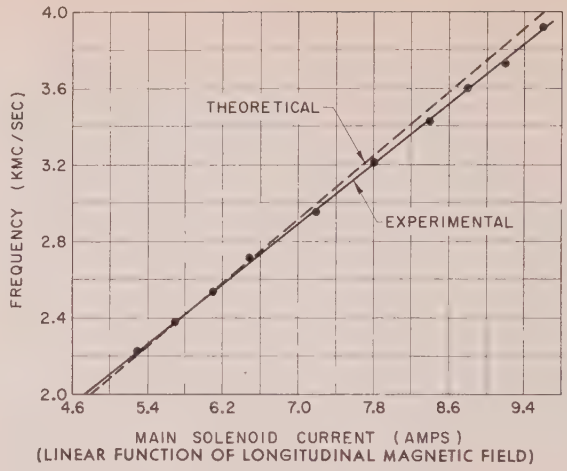

Fig. 4—Tuning chart showing the experimentally measured curve and the theoretical curve.

B. RF Power Output

RF power output has been measured over a frequency band from 2.1 to 3.9 kmc at different beam voltages and under varied conditions. Fig. 5 shows the peak power output vs frequency of oscillation for beam voltages of 3.0, 3.5 and 6 kv. Curves III and IV were taken at 3.5 and 3.0 kv, respectively, with a resistive load coupled through a finger-section coupler to the collector section of the tube. The "wiggles" in these output power curves resemble the "wiggles" measured in the reflection coefficient of the couplers. Coupler mismatch is more important for this tube than for the usual backward-wave oscillator, because the synchronous frequency for forward-wave amplification is very close to that for backward-wave oscillation.[5] Therefore, a signal reflected at the gun end can be amplified and can produce constructive or destructive interference with the original signal. The dash-line curve (Curve II) taken at 3.5 kv was with an *E-H* tuner inserted between the coupler and the resistive load so as to match the collector end. The couplers were not designed to be well matched near the cutoff frequency of the waveguide mode, and so the improvement in the output power near cutoff is more marked.

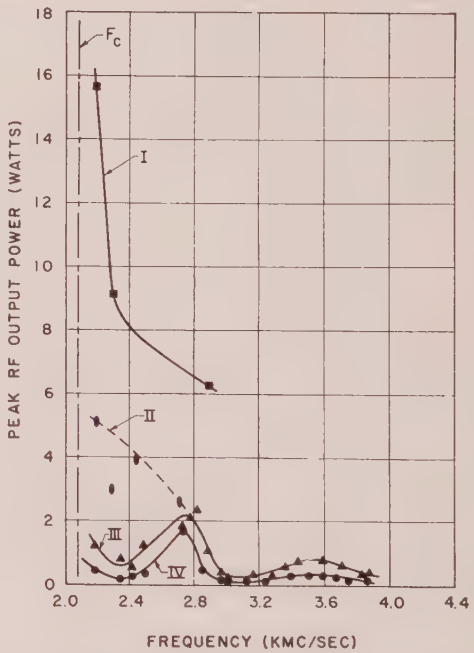

Fig. 5—RF power output vs frequency.
Curve I: Beam voltage = 6 kv. The collector end is terminated in an *E-H* tuner and then a resistive load.
Curve II: Beam voltage = 3.5 kv. Same termination as for Curve I.
Curve III: Beam voltage = 3.5 kv. The collector end is terminated in a resistive load without an *E-H* tuner.
Curve IV: Beam voltage = 3.0 kv. The collector end is terminated in a resistive load without an *E-H* tuner.
F_c = cutoff frequency of the waveguide mode.

The RF output pulse was very stable from about 2.1 to 2.9 kmc, and Fig. 6 is an oscillograph of the anode voltage pulse and the output pulse. The anode voltage is shown above the output signal; each division of the abscissa is 2 μsec, and it is seen that about one μsec is required for the oscillation to start. Above 2.9 kmc

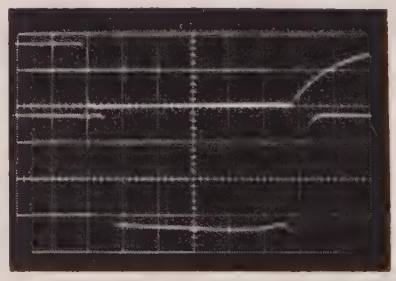

Fig. 6—Oscillograph showing anode voltage pulse as the top trace and RF output pulse as the bottom trace.

[5] R. H. Pantell, "Backward-wave oscillations in an unloaded guide," Proc. IRE, vol. 47, p. 1146; June, 1959.

the oscillation was less stable, and a more careful adjustment of the various parameters (such as compensation coil position) was necessary to obtain a steady output.

Since output power and efficiency (see section C below) have been observed to increase with beam voltage, it was thought desirable to increase the beam voltage as high as possible. After all measurements had been made, the beam voltage was increased to over 6 kv when breakdown inside the gun occurred. A series of measurements was then tried at 6 kv, shown in Fig. 5 as Curve I, giving 15.7 watts at 2185 mc and 9.14 watts at 2218 mc with an *E-H* tuner at the collector end. However, there were indications that the cathode and magnetic structure inside the gun had been damaged, and experiment was discontinued.

C. Tube Efficiency

Tube efficiency is measured over the same range of anode voltages as the power output and is defined as the ratio of RF output power to the dc beam power. The experimental results are shown in Fig. 7. Curves II, III, IV and V were taken at anode voltages of 3.5, 3.0, 2.5 and 2.0 kv, respectively, with a resistive load at the collector end. For the same reasons as those discussed in the RF power output section, there are "wiggles" in the curves and the efficiency is, in general, rather low. Matching the load at the collector end by inserting an *E-H* tuner greatly improves the efficiency near the cutoff frequency as shown by Curve I, taken at 3.5 kv. The resultant efficiency is greater than 1.5 per cent over 23 per cent of the frequency band near the cutoff frequency. Such an efficiency is comparable with the commercially available O-type backward-wave oscillators.

From Fig. 7, it is also clear that the tube efficiency increases with beam voltage on the range shown. No saturation efficiency had been reached in the beam-voltage range of 2.0–3.5 kv. Efficiency at 6 kv had also been measured and was somewhat less than that at 3.5 kv. However, since the gun was damaged, it is rather difficult to decide whether the saturation limit had been passed.

D. Beam Pitch Angle

Beam-pitch angle was determined by measuring the time delay between the collected current pulse and the emitted current pulse from the cathode. This time delay can be measured on a dual-beam oscilloscope, as shown in Fig. 8, which is an oscillograph of the leading edges of the pulses. The first trace is that of the emitted current, as seen through a pulse transformer; the second trace is that of the collected current as seen through another pulse transformer. The horizontal scale is 0.5 μsec/cm and the time delay can be seen clearly. However, when the tube was oscillating, the "ringing" in the pulse transformers became much worse and there was no means to identify corresponding points in the two traces exactly. Therefore, the beam-pitch angle so measured is subjected to perhaps a 50 per cent error.

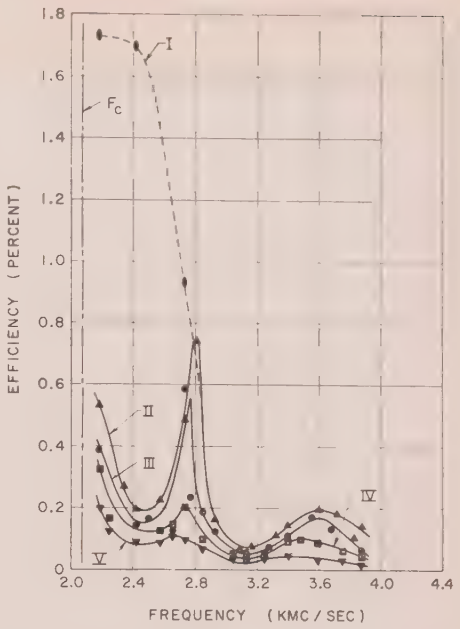

Fig. 7—Tube efficiency at various anode voltages.
Curve I: Beam voltage = 3.5 kv. The collector end is terminated in an *E-H* tuner and then a resistive load.
Curve II: Beam voltage = 3.5 kv.
Curve III: Beam voltage = 3.0 kv.
Curve IV: Beam voltage = 2.5 kv.
Curve V: Beam voltage = 2.0 kv.
Curves II–V were taken without the *E-H* tuner.

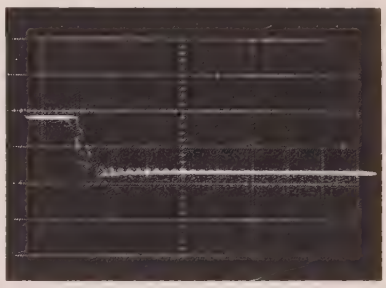

Fig. 8—Oscillograph showing time delay between emitted and collected currents.

The experimental results are shown in Fig. 9, for beam voltages of 3.0 kv and 3.5 kv, with auxiliary coil current and compensation coil position adjusted to give maximum oscillation power. It is seen that for 3-kv beam, the pitch angle is roughly constant at 9°, corresponding to a ratio of kinetic energy in the transverse motion to that in the longitudinal motion of about 36 to 1; and for 3.5-kv beam, the pitch angle remains constant at about 7°, corresponding to a ratio of transverse to longitudinal energies of approximately 67 to 1.

E. Harmonic Power Contents

Harmonic power is measured by passing the tube output through a high-pass filter and then into a very sensitive narrow-band receiver (the APR-9 radar receiver). By this means, harmonics up to the 4th, at 3.5 kv beam voltage, have been measured, and results are presented in Fig. 10. The second harmonic is about 60 db below the fundamental, at a power level of about −27 dbm over most of the frequency range. The third is roughly

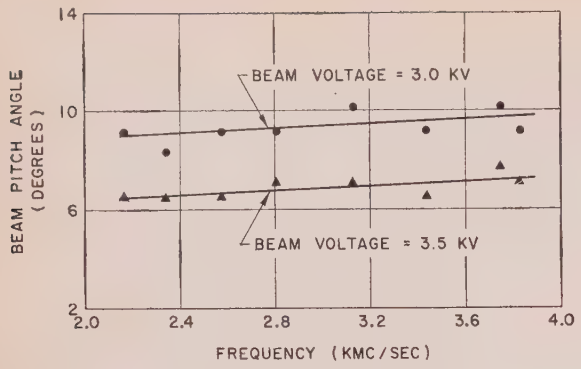

Fig. 9—Beam pitch angle vs frequency.

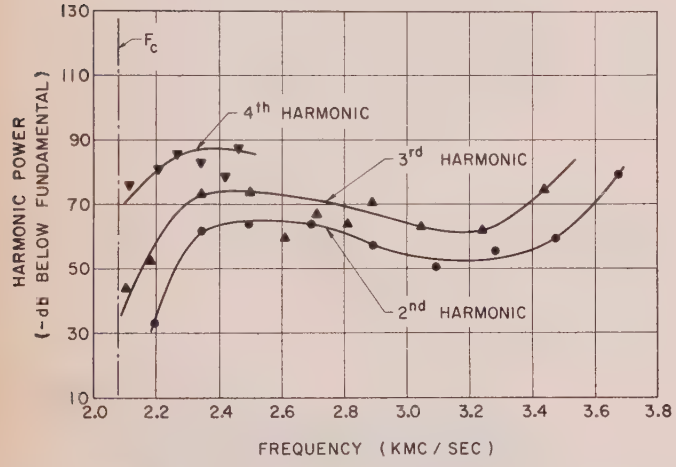

Fig. 10—Harmonic power contents of the tube.

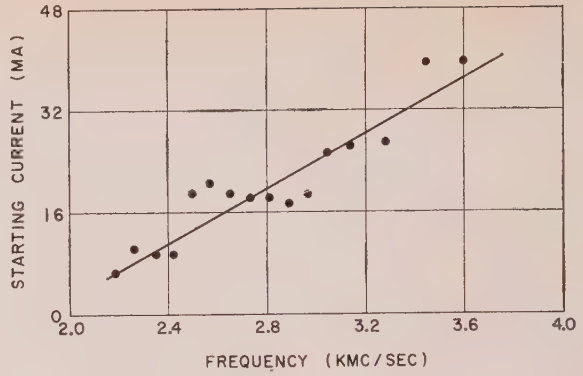

Fig. 11—Start-oscillation current of the tube as a function of frequency (beam voltage = 3.5 kv).

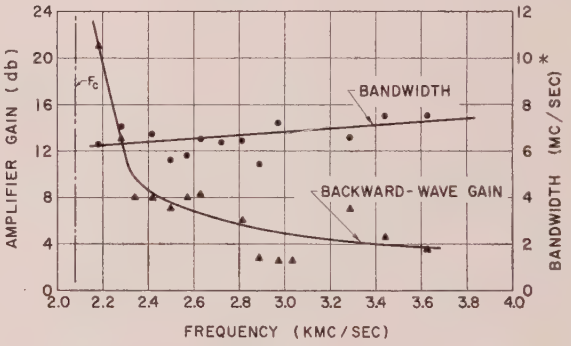

Fig. 12—Characteristics of the tube as a backward-wave amplifier (beam voltage = 3.5 kv). *Bandwidth is defined as the frequency band over which amplification occurs.

10 db below the second, and the fourth is about 10 db below the third.

F. Starting Current

Starting current was measured by adjusting the parameters of the tube for maximum oscillation power and then temperature-limiting the cathode. The results so obtained are shown in Fig. 11 for a beam voltage of 3.5 kv. The starting current has its lowest value near the waveguide mode cutoff frequency, and increases approximately linearly with frequency.

G. Backward-Wave Amplifier Characteristics

As a backward-wave amplifier, the tube has characteristics as shown in Fig. 12. Bandwidth was measured by keeping the dc flux density fixed and varying the signal frequency. Fig. 12 shows that the frequency band for which amplification occurred was extremely narrow, approximately 0.2 per cent. The backward-wave gain curve was obtained by adjusting the flux density for maximum amplification at each frequency. The measurements were made at 3.5 kv with the cathode temperature-limited. Backward-wave gain greater than 10 db occurs over approximately 10 per cent bandwidth near the cutoff frequency. The curves of Fig. 12 are with the beam current approximately 90 per cent of the starting current.

Conclusion

The experimental results presented in this paper indicate that it is possible to build a backward-wave oscillator and amplifier using the principle of cyclotron resonance interaction. The tube is tunable over a wide frequency range and does not require a slow-wave circuit. Near the cutoff frequency, the efficiency of this device is comparable to that of commercially available O-type oscillators. It is for the short millimeter wavelength region that this device will have its primary application. The dimensions of the structure are much larger than those of a helix operating at a similar beam voltage. For example, for operation in the two-to-one millimeter wavelength region, the transverse dimensions of the interacting circuit will be about 0.040 inch by 0.020 inch, and the RF power level should be in the tens-of-milliwatts range. Electron cyclotron resonance at one millimeter requires about 10^5 gauss, which can be obtained without too much difficulty for a millisecond pulse duration.

Acknowledgment

The authors wish to express sincere thanks to W. B. Lindsay and T. P. Lee for their helpful suggestions and collaboration, and to W. Bauer and P. Lee for their excellent work in constructing the tube.

Amplitude Scintillation of Radio Star at Ultra-High Frequency*

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Summary—Observations of radio star scintillation at a frequency of 915 Mc over a 12-month period (October, 1957–September, 1958) are described. By daily observations of Cygnus A (IAU 19N4A) over a wide range of elevation angles, both the amplitude and rate of scintillation were studied. It was found that the amplitude scintillation is strongest near the horizon and decreases rapidly as the altitude increases. The amount of the intensity fluctuation ranges from less than 3 per cent to over 50 per cent of the undisturbed intensity, while the rate of fluctuation varies from $\frac{1}{2}$ to 8 peaks per minute.

A fluctuation amplitude greater than 10 per cent was observed at low-altitude angles for 71 per cent of the total observations. The frequency of occurrence of scintillation follows a seasonal variation having two maxima; one in the winter and the other in the summer. During these two maximum periods, the mean fluctuation amplitude was found to be higher than during other periods. It was also found that the mean fluctuation rate in the winter is about three times faster than that in the summer.

It was also observed that the scintillation characteristics are markedly affected by the presence of auroras and/or geomagnetic disturbances. During the winter months, the fluctuation rate is roughly proportional to the three-hour geomagnetic K index.

I. Introduction

THE characteristics of radio waves passing through the earth's troposphere, ionosphere and exosphere have aroused considerable interest during the last several years. The characteristics of these radio waves bear directly on the seeing condition of radio stars and on the accuracy of tracking space vehicles and missiles. Accordingly, this problem is of great concern to radio astronomers, space communication engineers and antenna designers.

One of the phenomena associated with these radio waves is the fluctuations of its intensity and direction of arrival or "scintillation." The scintillation of radio stars was discovered by Hey, Parsons and Phillips[1] in 1946. While making a survey of the distribution of galactic radio radiation at 64 Mc, they found that the radio intensity received from the constellation of Cygnus exhibited peculiar fluctuations on certain occasions, whereas the intensity from the rest of the sky appeared to be constant with respect to time. They suggested that these fluctuations were probably caused by the existence of a few discrete radio sources of comparatively small angular dimensions with variable intensity.

This clue was followed up by Bolton and Stanley,[2] who discovered that the Cygnus region contained an intense radio source with an apparent angular diameter of less than eight minutes of arc. This radio source is now known as Cygnus A (IAU 19N4A). Further work by Bolton[3] resulted in the discovery of a number of other similar, though weaker, radio sources. Hundreds of such localized radio sources have since been detected by various workers and these sources are often referred to as "radio stars."

Until 1949, it was generally believed that the fluctuations of Cygnus A were inherent in the source itself and were perhaps similar to those found in sunspot radiation. However, Bolton found that the Cygnus A fluctuations observed simultaneously in Sydney, Australia, and in New Zealand were uncorrelated. This cast doubt on the variations being inherent in the source. Subsequent work by Smith[4] in Cambridge and Little and Lovell[5] in Manchester disclosed that there was no correlation between the fluctuations received at two points on the ground separated by a distance of 210 km, although fairly good correlation was obtained when the distance of the two receivers was reduced to 4 km. They concluded that the fluctuations are not inherent in the radio star itself, but are produced in the earth's ionosphere. More extensive work has been subsequently done by many groups, notably in England and in Australia. The results of this work have been recently summarized by Little, Rayton, and Roof,[6] and by Booker.[7] It should be pointed out that all these observations were made at frequencies below 100 Mc.

It is now believed that radio star scintillation is caused by irregularities in the earth's ionosphere. The irregularities have different RF refractive indexes and thus produce important effects on the propagation of radio waves. Therefore, when observed at ground level

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† Radio Observatory, Dept. of Elec. Engrg., The Ohio State University, Columbus Ohio.

[1] J. S. Hey, S. J. Parsons, and J. W. Phillips, "Fluctuations in cosmic radiation at radio-frequencies," *Nature*, vol. 158, p. 234; August 17, 1946.

[2] J. G. Bolton and G. J. Stanley, "Variable source of radio-frequency radiation in the constellation of Cygnus," *Nature*, vol. 161, pp. 312–313; February 28, 1948.

[3] J. G. Bolton, "Discrete sources of galactic radio-frequency noise," *Nature*, vol. 162, pp. 141–142; July 24, 1948.

[4] F. G. Smith, "Origin of the fluctuations in the intensity of radio waves from galactic sources," *Nature*, vol. 165, pp. 422–423; March 18, 1950.

[5] C. G. Little and A. C. B. Lovell, "Origin of the fluctuations in the intensity of radio waves from galactic sources," *Nature*, vol. 165, pp. 423–424; March 18, 1950.

[6] C. G. Little, W. M. Rayton, and R. B. Roof, "Review of ionospheric effects at VHF and UHF," PROC. IRE, vol. 44, pp. 992–1018; August, 1956.

[7] H. G. Booker, "The use of radio stars to sutdy irregular refraction of radio waves in the ionosphere," PROC. IRE, vol. 46, pp. 298–314; January, 1958.

on meter wavelengths, a radio star exhibits fluctuations in both intensity and position. The apparent fluctuations in intensity are called "amplitude scintillations," while the fluctuations in position are called "phase scintillations." These scintillations are analogous to the twinkling and dancing of optical stars at visual wavelengths.

Although radio star scintillations have been studied for some years, all observations were made at frequencies below 100 Mc. From the experimental evidence, as well as theoretical considerations of the ionosphere, it was generally believed that radio star scintillation would be negligibly small at frequencies above 300 Mc. In the past, there were very few attempts to detect scintillations at frequencies above 300 Mc. Australian workers (Bolton, Slee, and Stanley;[8] Bolton and Stanley[9]) made multifrequency observations of Cygnus A using sea interferometers in the frequency range of 40–300 Mc. They showed that the amplitude of scintillations decreases as the frequency increases up to 100 Mc, but remains nearly constant thereafter. The reason for the loss of frequency dependence at higher frequencies is not yet clear. They also reported that the occurrence of scintillation at frequencies above 200 Mc was very rare. Somewhat contrary to these results, Seeger[10] reported that at Ithaca, N. Y., the scintillation of Cygnus A at 205 Mc was always observed when the star's altitude angle is less than 20°, and on a few occasions the scintillation was observed at altitude angles of 60° to 90°. The reason for this discrepancy is not clear, but it is probable that it may be caused by difference in the properties of the upper atmosphere for the two sites, Sydney, Australia, and Ithaca, N. Y.

The interest in radio star scintillation at higher frequencies has increased during recent years, and several serious attempts have been made to study scintillation at frequencies above 300 Mc. Thus, scintillations were observed at 223 Mc and 456 Mc by Lansinger, *et al.*,[11] and at 915 Mc by the author.[12]

Furthermore, it was found that scintillations exist in the microwave region by Aarons, Barron, and Castelli[13] at 8-mm and 3-cm wavelengths, and by Kazés and Steinberg[14,15] at 3-cm wavelength. In both cases, radio emission from a localized region on the solar disk was used as the source. The scintillations in the microwave region are, however, most likely of tropospheric origin.

This paper describes an experimental investigation of radio star scintillation at a frequency of 915 Mc (33-cm wavelength) in which daily observations of Cygnus A were made over a wide range of altitude angles for one year (October, 1957–September, 1958) at The Ohio State University Radio Observatory. It was found that amplitude scintillation is strongest near the horizon and decreases rapidly as the altitude increases. The amplitude of intensity fluctuation ranges from less than three per cent to over 50 per cent of the undisturbed intensity, while the rate of fluctuation varies from $\frac{1}{2}$ to 8 peaks per minute.

A fluctuation amplitude greater than 10 per cent was observed at low-altitude angles for 71 per cent of the total observations. The frequency of occurrence of scintillation follows a seasonal variation having two maxima; one in the winter and the other in the summer. During these two maximum periods, the mean fluctuation amplitude was found to be higher than during other periods. It was also found that the mean fluctuation rate in the winter is about three times faster than that of the summer.

It has also been observed that the scintillation characteristics are markedly affected by the presence of auroras and/or geomagnetic disturbances. During the winter months, the fluctuation rate is roughly proportional to the three-hour geomagnetic K index.

II. Equipment and Observations

The radio star, Cygnus A (Right Ascension $19^h57^m45^s$; Declination $+40°35'$, Epoch 1950; IAU No. 19N4A), was chosen for the observations of scintillation for two reasons. Firstly, Cygnus A is the second strongest radio star in the sky and has a flux density of 22×10^{-24} watts meters^{-2} cps^{-1} at 915 Mc and it has an extremely small angular extent, of the order of two minutes of arc. Secondly, its altitude angle ranges from 0° to 90° (horizon to zenith) for observers at Columbus, Ohio (40°N latitude).

The equipment used for the scintillation observations consists of a 40-foot diameter, steerable parabolic antenna and a low-noise UHF radiometer. The antenna is equatorially mounted and the dish can be rotated from horizon to horizon so as to track Cygnus A from its rising to setting. The antenna feed is a half-wave dipole with circular ground plane. The dipole is tilted so that the response is horizontally polarized when the antenna

[8] J. G. Bolton, O. B. Slee, and G. J. Stanley, "Galactic radiation at radio frequencies: VI. Low altitude scintillations of the discrete sources," *Australian J. Phys.*, vol. 6, pp. 434–451; December, 1953.

[9] J. G. Bolton and G. J. Stanley, "Observations on the variable source of cosmic radio-frequency radiation in the constellation of Cygnus," *Australian J. Sci. Res. A*, vol. 1, pp. 58–69; March, 1948.

[10] C. L. Seeger, "Some observations of the variable 205 mc/s radiation of Cygnus A," *J. Geophys. Res.*, vol. 56, pp. 239–258; June, 1951.

[11] J. M. Lansinger, C. G. Little, R. P. Merritt, and E. Stiltner, "Fluctuations in the Apparent Amplitude and Position of Extra-Terrestrial Radio Sources, as Observed Near the Auroral Zone," presented at IRE-URSI Symp., Washington, D. C.; May 22–25, 1957.

[12] H. C. Ko, "Amplitude scintillation of extraterrestrial radio waves at ultra-high frequency," Proc. IRE, vol. 46, pp. 1872–1873; November, 1958.

[13] J. Aarons, W. R. Barron, and J. P. Castelli, "Radio astronomy measurements at VHF and microwaves," Proc. IRE, vol. 46, pp. 325–333; January, 1958.

[14] I. Kazés, "Étude de la scintillation du soleil observée sur la longueur d'onde de 3.2 cm," *Compt. rend. acad. sci.*, vol. 245, pp. 635–639; August 5, 1957.

[15] I. Kazés and J. L. Steinberg, "Étude de la scintillation du soleil observée avec plusiers antennes sur la longueur d'onde de 3.2 cm," *Compt. rend. acad. sci.*, vol. 245, pp. 782–785; August 12, 1957.

is pointed toward the western horizon, but is vertically polarized when pointed toward the eastern horizon. At 915 Mc, the antenna has a pencil beam of 1.7° between the half-power points and directivity of 39 db with respect to an isotropic radiator. The effective collecting aperture to the antenna is approximately 75 square meters.

The radiometer is a total-power type, crystal superheterodyne receiver. It consists of a crystal mixer input followed by a low-noise 30-Mc IF preamplifier, postamplifier and second detector. The IF amplifiers have a bandwidth of 2.8 Mc. The noise figure of the receiver is about 7 db. The output of the second detector is followed by an automatic pen recorder. The chart speed was 0.4 inch per minute. An output time constant of two seconds was used in the observations.

With the antenna pointing towards the sky, the antenna temperature, or the effective temperature of the antenna radiation resistance, is given by

$$T_a = \frac{1}{4\pi}\int_{4\pi} T_s(\theta,\Phi)G(\theta,\Phi)d\Omega$$

where $T_s(\theta,\Phi)$ is the effective temperature of the sky and $G(\theta,\Phi)$ is the gain function of the antenna. At the transit of Cygnus A across the antenna beam, the antenna temperature is increased by about 70°K. The rms fluctuations in the output of the recorder due to the receiver noise correspond to an antenna temperature variation of about 1°K, thus giving an "on-the-record" SNR of 70 or 18.5 db.

Observations of the radio star, Cygnus A, were carried out over one year from October, 1957 to September, 1958 at The Ohio State University Radio Observatory. The observatory is located at longitude 83°03′ west and latitude 40°01′ north. Geomagnetic coordinates of the observatory are geomagnetic longitude 342°.8 east and geomagnetic latitude 51.3° north. At this location, Cygnus A rises from approximately 32° east of north, reaches the zenith at its meridian transit, and sets at 32° west of north. The radio star is above the horizon for about 18 hours a day. Fig. 1 shows the change of the elevation and azimuth angles of Cygnus A during one day.

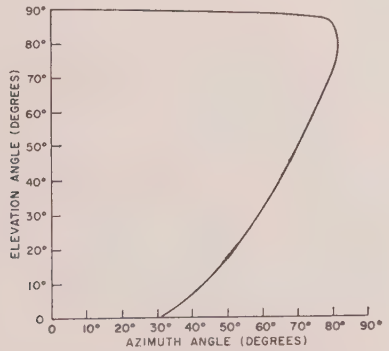


Fig. 1—The change in the elevation and azimuth angles of Cygnus A during one day at 40°N latitude.

The observations were made by setting the antenna beam slightly west of Cygnus A in hour angle (about 3.7°), and the drift curve recorded as the radio star passed through the antenna beam. The procedure was repeated at intervals of 7.5° in hour angle. Thus, 18 drift curves could be obtained as the star changed its altitude angle from 0° to 90° or 90° to 0°. Fig. 2 shows some of the drift curves recorded on the night of November 24–25, 1957.

During the period from October, 1957–September, 1958, 271 series of observations were made of Cygnus A at its setting, and 77 series of observations at its rising. Observations at setting were made throughout the year, but at rising, only during the summer. Fewer records were obtained at its rising because of frequent man-made radio interference in the northeast direction.

Because of the difference between the sidereal time and mean solar time, the time of day at which a star rises above the horizon advances four minutes each day and moves through 24 hours during the course of a year. Thus, in January, Cygnus A rises early in the morning and sets late in the evening, while in July, it rises late in the afternoon and sets late in the morning.

III. Analysis of Data

The observed drift curves of Cygnus A showed various forms of intensity fluctuations. The general character of the amplitude scintillations may be described by the magnitude of the intensity fluctuations and the rate at which these fluctuations occur. In the analysis of records, it is convenient to assign a quantitative measure to the character of the scintillations. Thus, a mean value of fluctuation amplitude and a fluctuation rate were assigned to each drift curve. The *fluctuation amplitude* was measured by the ratio of the mean peak-to-trough variations in the intensity to its mean intensity, and is expressed as a percentage. The *fluctuation rate* was determined by the average number of peaks per minute. Fig. 3 shows samples of scintillation records obtained at altitude angle of 3.7° on different days. The records illustrate quiet [Fig. 3(a)], weak [Fig. 3(b)], intermediate [Fig. 3(c)], and highly disturbed [Fig. 3(d)] conditions. The fluctuation amplitude and rate for each record are also shown. The observed scintillation data (October, 1957 to September, 1958) were analyzed in terms of:

A) the altitude dependence of scintillation,
B) the frequency of occurrence of scintillation,
C) the seasonal or diurnal variations of the mean fluctuation amplitude and mean fluctuation rate,
D) the correlation between the fluctuation amplitude and fluctuation rate.

The observations made at the setting and at the rising were analyzed separately. The effects of geomagnetic storms and auroras upon the radio scintillations will be described in Section IV.

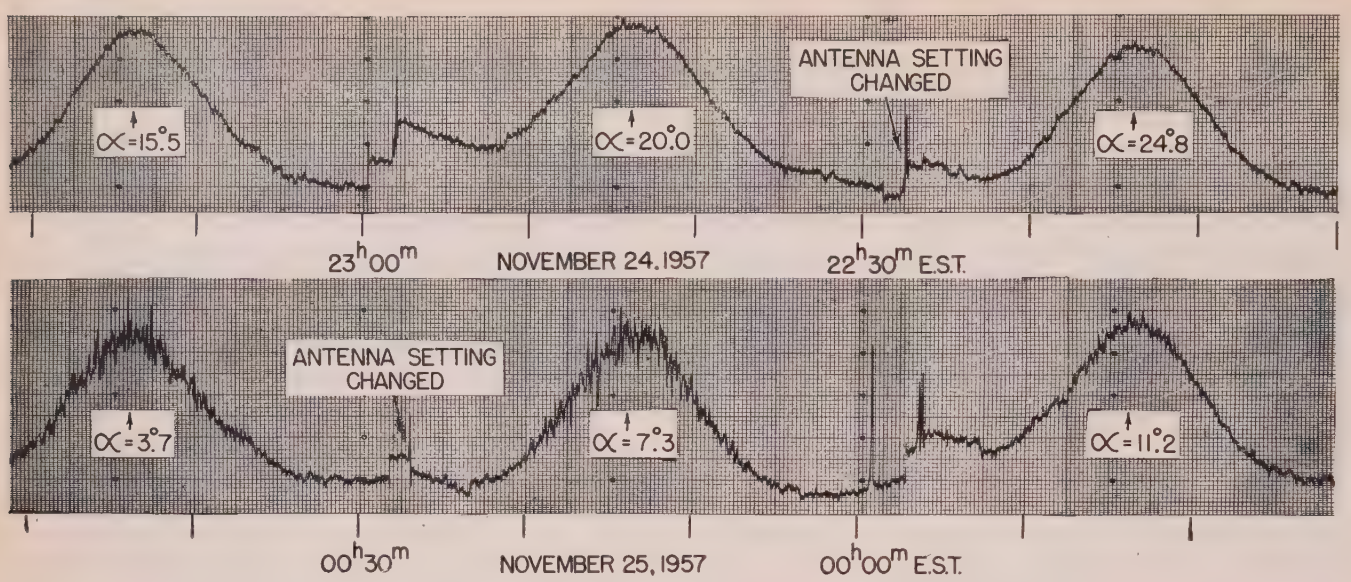


Fig. 2—Radio scintillation records of Cygnus A at its setting on November 24–25, 1957, taken with a 40-foot diameter, parabolic reflector antenna at 915 Mc.

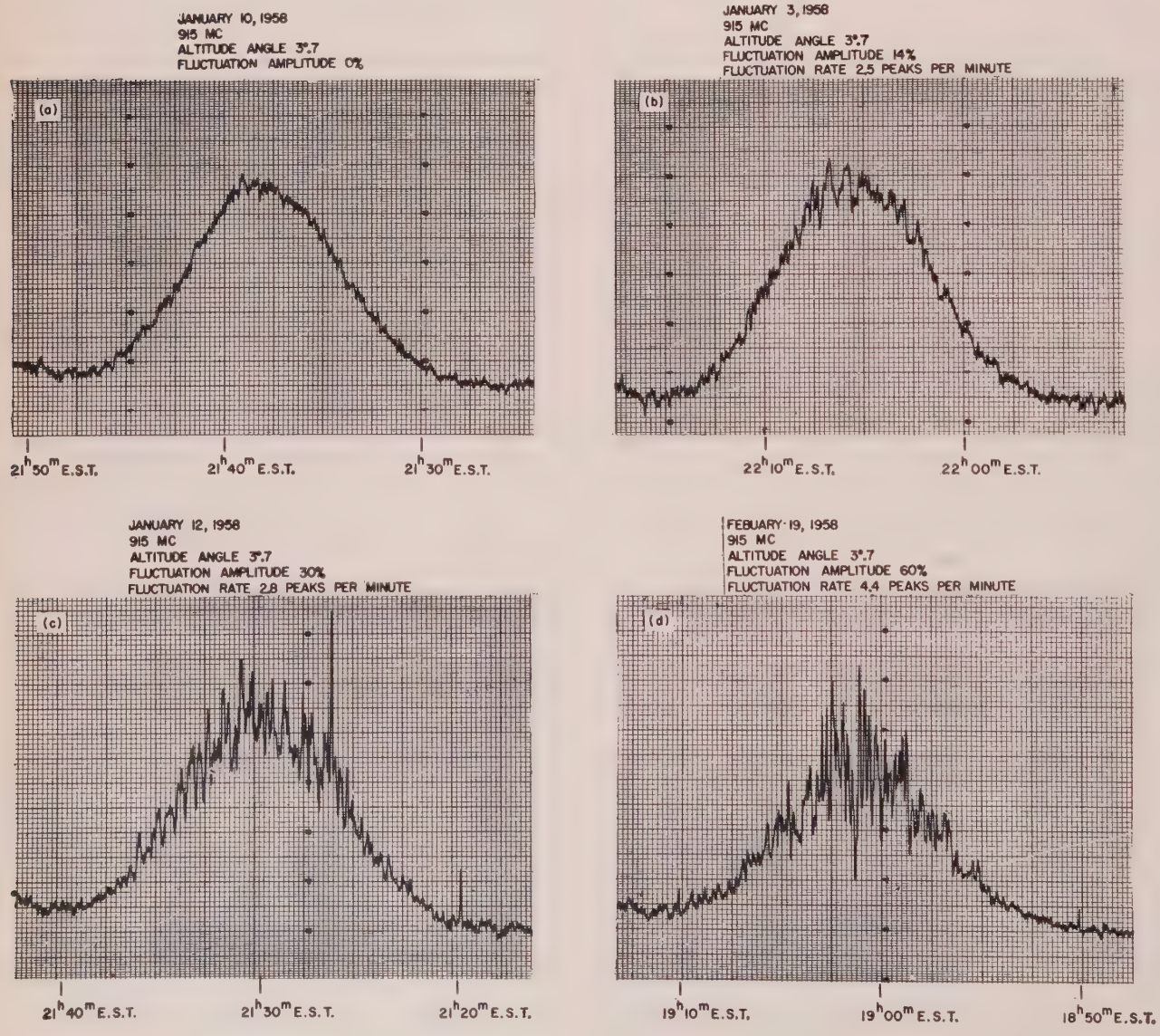


Fig. 3—Sample drift curves of Cygnus A showing various degrees of amplitude scintillations. The records illustrate (a) quiet, (b) weak, (c) intermediate, and (d) highly disturbed conditions.

A. Dependence of Scintillation on the Altitude of the Radio Star

Over 95 per cent of the observed records showed that the amplitude of scintillation is strongest near the horizon and decreases rapidly as the altitude angle of Cygnus A is increased. At altitude angles above 15°, the scintillation was either too small to be detected (less than three per cent) or did not exist in these records. The manner in which the fluctuation amplitude diminishes with the altitude angle of the radio star varies from day to day as shown in Fig. 4.

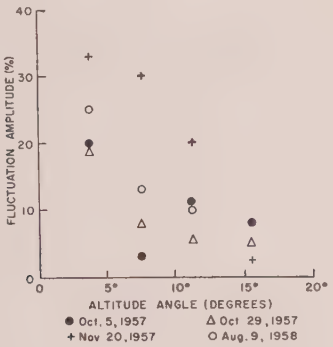


Fig. 4—Fluctuation amplitude as a function of altitude angle.

The diminution of scintillation with increasing altitude angle is further shown in Fig. 5. In this figure, the percentage of days throughout the year on which fluctuation amplitudes greater than 10 per cent were observed is plotted as a function of the altitude angle. Fig. 5 shows that, as the altitude angle of the radio star increases, the probability of observing scintillation decreases rapidly.

On a few rare occasions, the scintillations were observed not only at low angles of elevation, but also at high angles of elevation up to 90°. Furthermore, the fluctuation amplitude at high-altitude angles was greater than that at low altitudes. On these occasions, visible auroras were also observed near overhead.

The fluctuation rate was also found to vary with the altitude angle. As the altitude angle increases, the fluctuation rate may increase, decrease, or remain nearly constant. Typical data are illustrated for each case in Fig. 6. The cases shown in Fig. 6(a) and (c) (*i.e.*, rate increasing or constant) are much more frequently observed than the case in Fig. 6(b) (*i.e.*, rate decreasing).

B. Frequency of Occurrence of Scintillation

Observations of the setting of Cygnus A throughout the year at altitudes below 5° indicated that the fluctuation amplitude was greater than 10 per cent on 194 days out of 271 days, or for 71 per cent of the observations, and the fluctuation amplitude was greater than 20 per cent on 77 days out of 271 days, or for 28 per cent of the observations. The frequency of occurrence of scintillation, however, was observed to vary during the course of

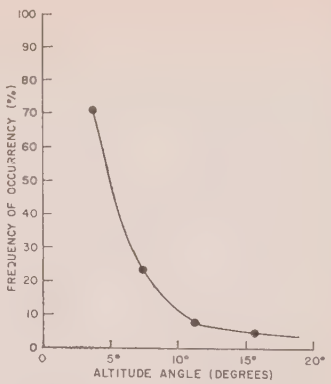


Fig. 5—Frequency of occurrence of scintillation as a function of altitude angle.

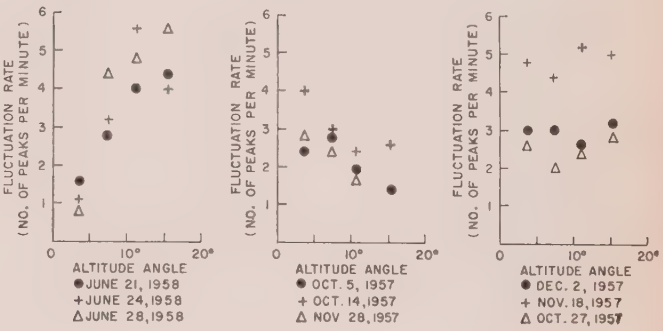


Fig. 6—Fluctuation rate as a function of altitude angle.

a year. To show this, the percentage of days during each month on which the fluctuation amplitude was greater than 10 per cent and 20 per cent is plotted in Fig. 7 for the setting and rising of Cygnus A, respectively. The figures show the probability that at altitude angles less than 5°, the fluctuation amplitude would be greater than a certain magnitude (*i.e.*, 10 per cent and 20 per cent) for each month.

There are two well-defined maxima in both Fig. 7(a) and (b), one in the winter (November and December) and the other in the summer (July and August). There is a deep minimum in April and a small minimum in September. Since the solar time of the setting (or the rising) of Cygnus A varies over 24 hours throughout the year, these observed seasonal variations could be due to a purely diurnal variation. Accordingly, the frequency of occurrence of scintillation was also plotted as a function of local solar time and is shown in Fig. 8.

It should be noted that the seasonal variation of the frequency of occurrence of scintillation for the setting of Cygnus A is similar to the one for the rising (see Fig. 7). On the other hand, Fig. 8 shows that the diurnal variation of the frequency of occurrence of scintillation for the setting is quite different from the one for the rising. This indicates that the variation of the frequency of occurrence of scintillation is largely seasonal, and the diurnal variation, if present, is probably small. However, in the present series of observations, it is not possible to separate the seasonal and diurnal effects.

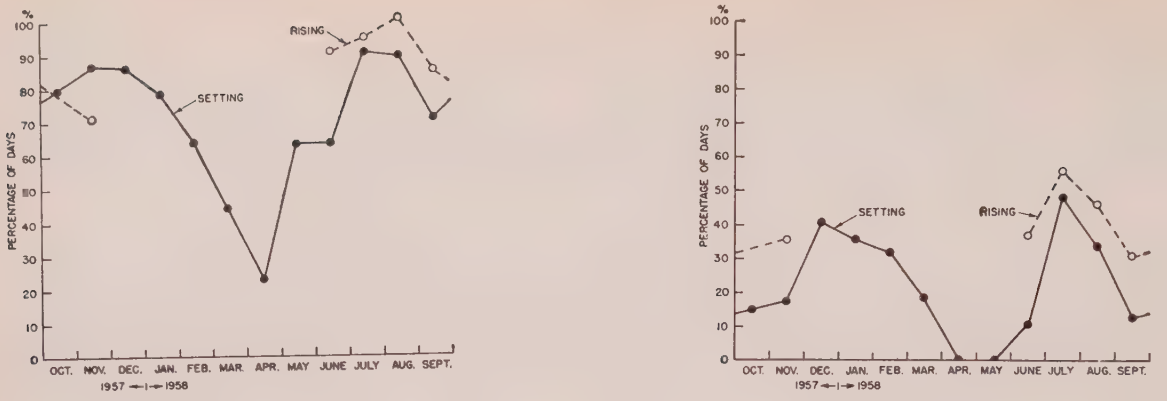


Fig. 7—(a) Frequency of occurrence of scintillation showing the percentage of days during each month on which the fluctuation amplitude is greater than 10 per cent. (b) Frequency of occurrence of scintillation showing the percentage of days during each month on which the fluctuation amplitude is greater than 20 per cent.

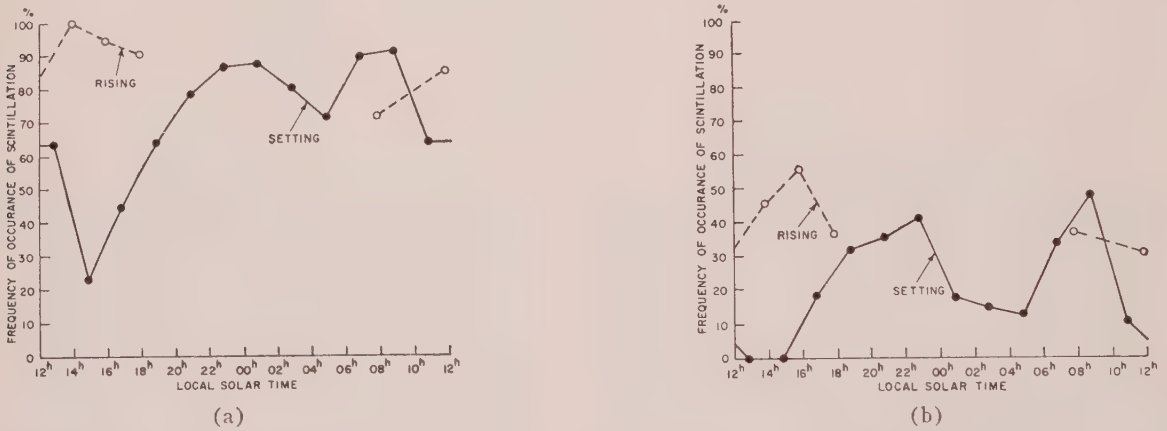


Fig. 8—(a) Frequency of occurrence of scintillation as a function of local solar time for fluctuation amplitude greater than 10 per cent. (b) Frequency of occurrence of scintillation as a function of local solar time for fluctuation amplitude greater than 20 per cent.

C. Seasonal Variation of Monthly Mean Fluctuation Amplitude and Fluctuation Rate

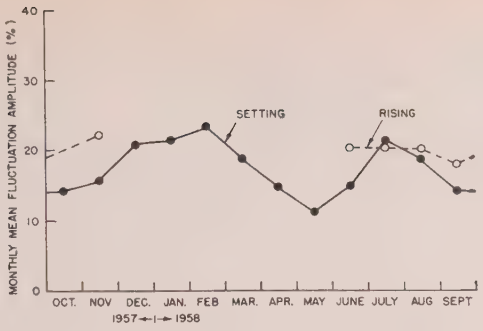
It has been shown in the previous section that the frequency of occurrence of scintillation has a seasonal variation (see Fig. 7). Accordingly, a study was made of the variation of the monthly mean values of both fluctuation amplitude and fluctuation rate. Only the records which showed fluctuation amplitude greater than 10 per cent at low-altitude angle were used in the analysis. The data consist of 194 records obtained at the setting of Cygnus A and 60 records at its rising. It should be pointed out that for the months in which frequency of occurrence of scintillation is low (*i.e.*, March and April), the number of records was small. Accordingly, the statistics of the mean values were poor for these months.

The monthly mean values of fluctuation amplitude and fluctuation rate are plotted for each month in Fig. 9. Fig. 9(a) shows that the variation of the monthly mean fluctuation amplitude has two well-defined maxima, one in the winter and the other in the summer. These maxima coincide roughly with those found in the seasonal variation of the frequency of occurrence of scintillation (see Fig. 7). Thus the high frequency of occurrence of scintillation in the winter and summer appears to be associated with the large mean fluctuation amplitude.

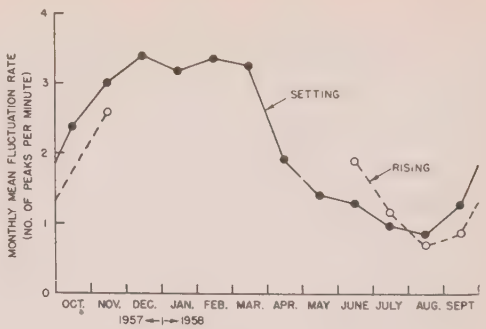
Fig. 9(b) shows that the monthly mean fluctuation rate has a single maximum throughout the year. The mean fluctuation rate during the winter period is about three times faster than during the summer months. A slow type of fluctuation is common during the winter, while a fast fluctuation is common during the summer, as illustrated in Fig. 10.

On some occasions, a combination of the two types of fluctuations was observed—one having a high rate and small amplitude of fluctuation, and superimposed on another having a low rate of fluctuation and a large fluctuation amplitude. This type of scintillation was, however, very rare and was observed only during the summer.

Since the local solar time of the setting or the rising of Cygnus A varies over 24 hours throughout the year, the seasonal variation could be due to a purely diurnal variation. Accordingly, the mean values of the fluctuation amplitude and fluctuation rate are also plotted as a function of local solar time and are presented in Fig. 11.



(a)


MONTHLY MEAN FLUCTUATION RATE (NO. OF PEAKS PER MINUTE)
SETTING
RISING
OCT. NOV. DEC. JAN. FEB. MAR. APR. MAY JUNE JULY AUG. SEPT
1957 1958

(b)

Fig. 9—(a) Seasonal variation of the monthly mean fluctuation amplitude. (b) Seasonal variation of the monthly mean fluctuation rate.

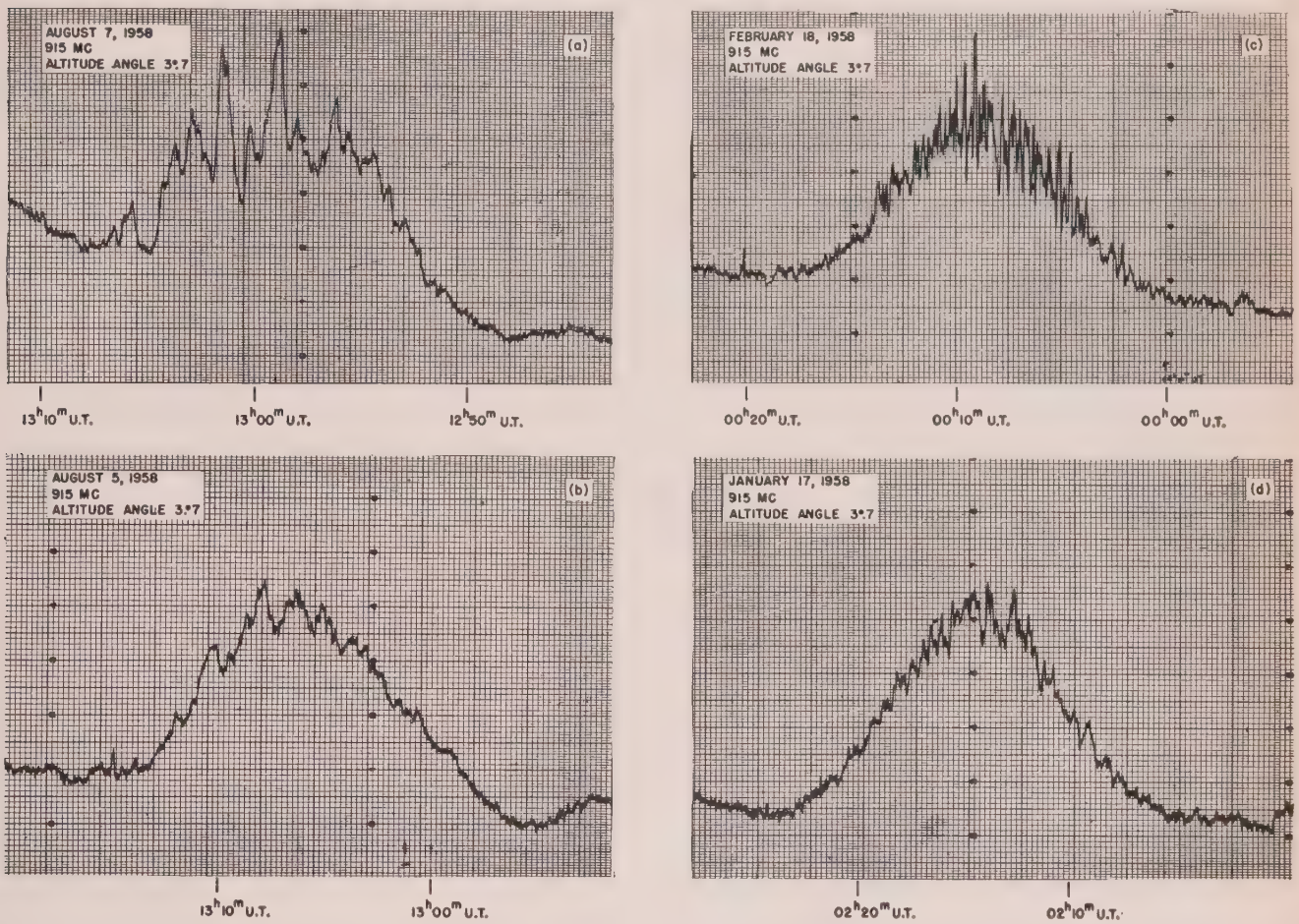
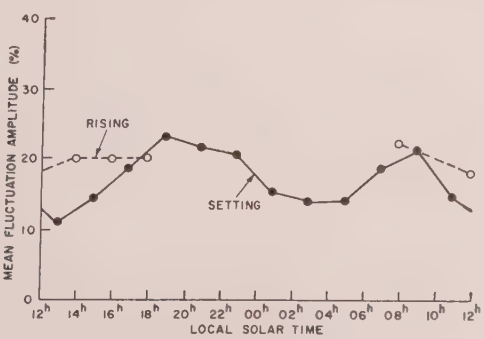
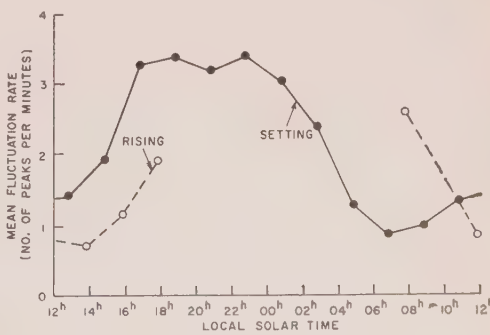


Fig. 10—Typical scintillation records of Cygnus A. (a) Strong scintillation during the summer. (b) Weak scintillation during the summer. (c) Strong scintillation during the winter. (d) Weak scintillation during the winter. Note that the scintillation rate in the summer records (a) and (b) is smaller than that in the winter records (c) and (d).



(a)


MEAN FLUCTUATION RATE (NO. OF PEAKS PER MINUTES)
RISING
SETTING
12h 14h 16h 18h 20h 22h 00h 02h 04h 06h 08h 10h 12h
LOCAL SOLAR TIME

(b)

Fig. 11—(a) Mean fluctuation amplitude vs local solar time. (b) Mean fluctuation rate vs local solar time.

Comparison of Fig. 9 and Fig. 11 shows that the variations of the mean fluctuation amplitude and mean fluctuation rate are seasonal variations, and the diurnal variation, if present, is probably small. Fig. 11 shows that the fluctuation amplitude is higher shortly before midnight and noon, while the fluctuation rate is higher near midnight.

D. *Fluctuation Amplitude vs Fluctuation Rate*

The altitude dependence of the fluctuation amplitude and the fluctuation rate has been discussed in a previous section (see Figs. 4–6). A study of any correlation between the fluctuation amplitude and fluctuation rate must be made at a fixed altitude angle in order to eliminate the altitude effect. Thus, over 240 observations of Cygnus A at its setting (at altitude of 3.7°) were analyzed. It was found that there was in general no correlation between the fluctuation amplitude and rate, except during the months of March and September. The scatter diagram of the fluctuation amplitude and fluctuation rate for March shows some positive correlation between the two, while the diagram for September shows a slight negative correlation.[16]

IV. Radio Star Scintillation During Auroras and Geomagnetic Disturbances

The effects of auroras and magnetic storms on radio star scintillation were first noted by Little and Maxwell.[17] While studying the scintillation of the radio stars Cassiopeia A and Cygnus A at 81.5 mc, Little and Maxwell found that, during auroras, the fluctuation rate is on the average four times more rapid than during normal conditions. However, there was no one-to-one correlation between the fluctuation amplitude and the intensity of auroras. They also noted that the fluctuation rate is approximately proportional to the geomagnetic K-index. Hewish[18] and Dueno[19] have also found a similar correlation between the fluctuation rate and K-index. A more detailed investigation of this relation has been made recently by Dagg.[20,21] Dagg found that, although there is no one-to-one correlation between the fluctuation amplitude and the geomagnetic activity, there is a correlation between the two over an extended period of time. It should be pointed out here that all the relations described above were found at frequencies below 100 Mc.

During the course of present observations at 915 Mc, it was found that auroras and/or geomagnetic disturbances also showed marked effects on the scintillation characteristics. Since the observations of Cygnus A at its setting during the period from September, 1957 to February, 1958 were made in the dark hours (18^h00^m to 04^h00^m local solar time), it was possible to compare these scintillation records with visual auroral observations. The U. S. Visual Auroral Observations published by the Aurora Data Center, Cornell University, Ithaca, N. Y., were used for comparison with the radio scintillation data. The scintillation records obtained during 28 auroras were compared with those obtained in non-auroral periods. It was found that during the auroral displays, the fluctuation rate increases by a factor of about 1.5, and on some occasions the fluctuation amplitude was also found to increase slightly.

On a few rare occasions, the auroras were so intense as to be visible overhead and lasted for a long period of time. During these intense auroras, the scintillations were observed not only at low altitudes but also at altitudes near 90°. The scintillation observed at high altitudes is caused by the passing of radio waves from Cygnus A through the zenithal auroral display. On September 23, 1957, a strong aurora was observed at Columbus, which lasted for the unusually long period of 10 hours. The aurora was one of the most intense recorded in 1957. The aurora commenced at 01^h00^m UT and reached its maximum at about 05^h00^m UT, at which time the aurora was visible overhead. The radio scintillation record on this date is shown in Fig. 12. The radio observation was started at 01^h00^m UT when Cygnus A was near the zenith. Strong radio scintillation commenced at 04^h40^m UT when the altitude of Cygnus A was about 50°. The scintillation activity lasted until 10^h00^m UT when Cygnus A set.

Another strong aurora was observed on September 24, 1957. The aurora commenced at 03^h00^m UT and lasted until 06^h00^m UT. The radio scintillation was first observed at about 02^h10^m UT and lasted until 05^h00^m UT. During this period, the altitude angles of Cygnus A were between 80° and 50°. After 05^h00^m UT, no scintillation was observed until 09^h00^m UT when the altitude angle of Cygnus A was about 8°.

Auroras are almost always accompanied by severe geomagnetic disturbances. Accordingly, it is of interest to compare the radio scintillation characteristics with the three-hour K-index of geomagnetic activity during both disturbed and nondisturbed periods. In Fig. 13 are shown scatter diagrams of fluctuation rate vs three-hour geomagnetic K-index for each month. All the data refer to the observations of Cygnus A at the altitude of 3.7° at its setting. The three-hour geomagnetic K-indexes were supplied by the Meanook Geomagnetic Observatory, Meanook, Alberta, Canada. Fig. 13 shows that during the months of November, December, Janu-

[16] H. C. Ko, "Radio star scintillation at 915 Mc," The Ohio State University, Columbus, Radio Observatory Rept. No. 12; April, 1959.

[17] C. G. Little and A. Maxwell, "Scintillation of radio stars during aurorae and magnetic storms," *J. Atmos Terres. Phys.*, vol. 2, no. 6, pp. 356–360; 1952.

[18] A. Hewish, "The diffraction of galactic radio waves as a method of investigating the irregular structure of the ionosphere," *Proc. Roy. Soc. A*, vol. 214, pp. 494–514; October 9, 1952.

[19] B. Dueno, "Low-angle fluctuations of the radio star Cassiopeia as observed at Ithaca, N. Y., and its relation to the incidence of sporadic-E," *J. Geophys. Res.*, vol. 61, pp. 535–540; September, 1956.

[20] M. Dagg, "The correlation of radio star scintillation phenomena with geomagnetic disturbances and the mechanisn of motion of the ionospheric irregularities in the F region," *J. Atmos. Terres. Phys.*, vol. 10, no. 4, pp. 194–203; 1957.

[21] M. Dagg, "Diurnal variations of radio star scintillations, spread F and geomagnetic activity," *J. Atmos. Terres. Phys.*, vol. 10, no. 4, pp. 204–214; 1957.

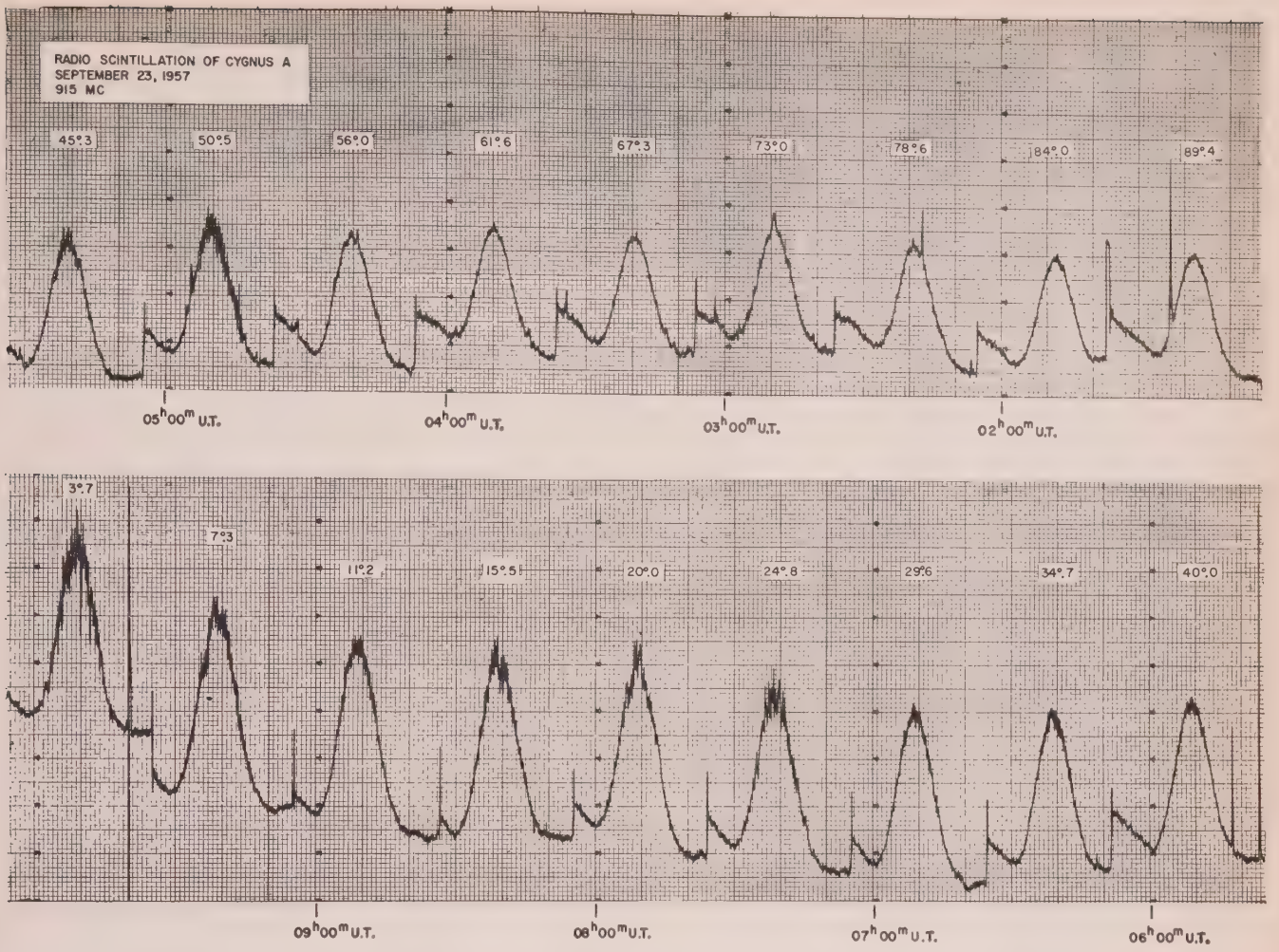


Fig. 12—Records of Cygnus A obtained on September 23, 1957, showing scintillation during a prolonged aurora. The numbers above the drift curves give the altitude angle of Cygnus A at the time of observation.

ary, and February, the fluctuation rate increases as the K-index increases. On the other hand, for the other months, the correlation between the fluctuation rate and the K-index is very poor. It should be also noticed that during the winter months the mean fluctuation rate is on the average three times faster than that for the summer months [also see Fig. 9(b)]. This may indicate that the fast scintillations observed in the winter which correlate well with the K-index, and the slow scintillations observed in the summer which correlate poorly with the K-index, may be caused by different mechanisms.

The fluctuation amplitude observed at the altitude of 3.7° at setting of Cygnus A was also compared with the three-hour geomagnetic K-index. In general, the correlation between the two is poor.[16]

V. Discussion

A comparison of the above results with the scintillation characteristics observed at meter wavelengths indicates that some features are very similar, while others are quite different. The results of the comparison are as follows:

1) The fluctuation rate observed at 915 Mc is about $\frac{1}{2}$ to 8 peaks per minute, which is the same order of magnitude as that observed at meter wavelengths.
2) The average fluctuation amplitude observed near the horizon at 915 Mc is about 15 per cent. This is about one-sixth that observed at 81.5 Mc.[22]
3) The scintillation decreases rapidly as the elevation angle increases both at 915 Mc and at meter wavelengths.
4) At 915 Mc, the scintillation shows strong seasonal variation and little diurnal variation. At meter wavelengths, English observers[6] found only diurnal variation, while Australian observers[8] found both seasonal and diurnal variation. At 915 Mc, it was also found that the fluctuation rate shows a strong seasonal variation, the rate during the winter months being approximately three times greater than that for the other months. However, no such seasonal variation was reported by English[6] or Australian[8] observers.

[22] C. G. Little and A. Maxwell, "Fluctuations in the intensity of radio waves from galactic sources," *Phil. Mag.*, vol. 42, pp. 267–278; March, 1951.

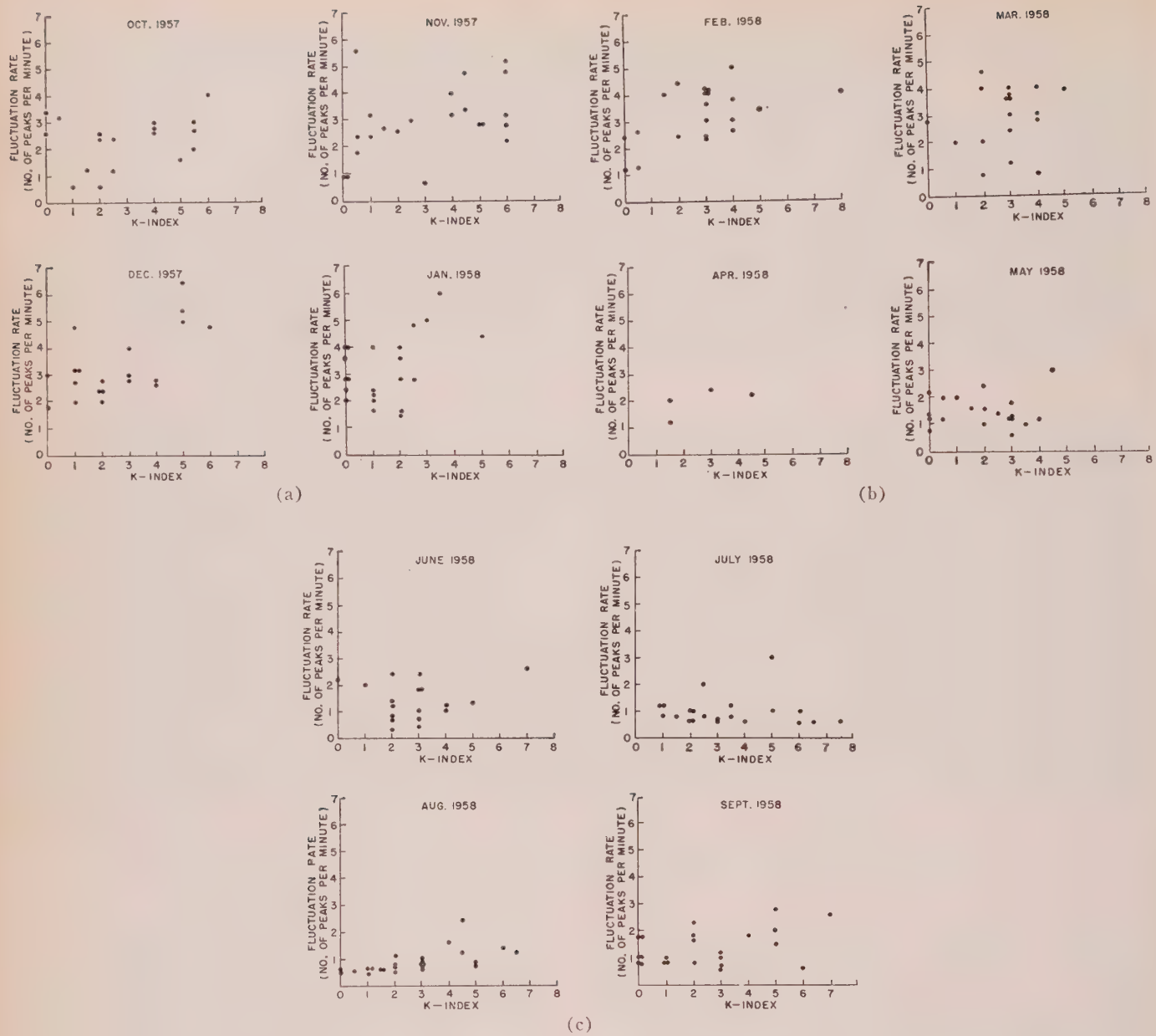



Fig. 13—Fluctuation rate vs 3-hour geomagnetic K-index.

5) The scintillation is markedly affected by the presence of auroras, at 915 Mc as well as at meter wavelengths.
6) The fluctuation rate increases with the geomagnetic K-index both at 915 Mc and at meter wavelengths. However, at 915 Mc it was found that such correlation exists only during the winter months. On the other hand, observations by Little and Maxwell[17] at 81.5 Mc indicate no such seasonal dependence, while observations by Dueno[19] at 52 Mc show that such correlation exists during the spring, summer, and autumn months but ceases during the winter months.

It is generally agreed that the scintillation observed at meter wavelengths is of ionospheric origin. The results of the present investigation of radio scintillation at 915 Mc suggest that the scintillation at 915 Mc is also probably of ionospheric origin. This view is supported by the facts that the scintillation at 915 Mc is markedly affected by the presence of auroras and geomagnetic disturbances, and some of the general features of scintillation at 915 Mc are very similar to those observed at meter wavelengths. However, it is also probable that part of the scintillations observed at 915 Mc may be of tropospheric origin. Further experiments are needed to assess and separate the origins of scintillation at 915 Mc.

VI. Acknowledgment

The author is indebted to Dr. J. D. Kraus for helpful discussions, and to K. W. Olson and C. E. Shuster for assistance in the maintenance of the equipment and the taking of observations.

Voltage Breakdown of Antennas at High Altitude*

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Summary—**The factors influencing the power-handling capability of antennas at high altitude are considered in this paper. The physical mechanism involved, including the roles of attachment, free diffusion, ambipolar diffusion, and nonuniform field distribution in the breakdown process, is qualitatively described. These factors are illustrated by breakdown curves for various antenna configurations under both CW and pulse conditions. Normalized data—useful for estimating breakdown fields when the conditions for scaling are fulfilled—are presented.**

The effect of missile environment on breakdown characteristics is discussed, and an experiment that involves artificially introducing ionization near the surface of the antenna is described.

Methods are then considered for increasing the power-handling capability, and typical results are given showing the increase in power that can be achieved.

I. Introduction

EXAMINATION of the received signal from electronic systems on various missiles has indicated the presence of antenna voltage breakdown in the altitude range of 50,000 to 300,000 feet at VHF, UHF and microwave frequencies. The presence of antenna voltage breakdown is manifested by a decrease in transmitted signal intensity, change in the input impedance and radiation pattern of the missile antenna, modification of the pulse shape, and noise modulation on the signal.[1] Since it is important, in many cases, to obtain continuous telemetry and tracking information to determine impact position and system performance data, an experimental investigation has been carried out to study the power handling capabilities of antennas at high altitude. The principal aims of this study are 1) to determine the physical mechanism involved, 2) to measure and tabulate experimental data on breakdown field for various types of antenna, and 3) to study methods for improving the power handling capability of antennas.

II. Breakdown Mechanism

The high frequency breakdown mechanism has been thoroughly investigated in the past for circuit elements such as transmission lines and cavities.[2,3] The source of primary ionization is due to electron motion, and breakdown occurs when the gain in electron density becomes equal to the loss of electrons by diffusion and attachment. This relationship is expressed as:

$$\frac{\partial n}{\partial t} = (\nu_i - \nu_a)n + S + \nabla^2(Dn) \tag{1}$$

where

n is the electron density (electrons/cc)
ν_i is the ionization rate (ionizations/sec/electron)
ν_a is the attachment rate (attachments/sec/electron)
S is the ionization produced by an external source (electrons/cc sec)
D is the diffusion coefficient for electrons (cm²/sec).

An examination of the factors making up (1) shows that, for geometrically similar antennas, frequency scaling is applicable.[4] Also, if the form of the near-zone field remains the same with some varying dimension (as with parallel plate-transmission lines with varying spacing, narrow slot antennas with varying width, and monopoles with varying diameter), the value of the ratio of effective electric field required for breakdown, E_e, to pressure, p, is a function of pressure times the characteristic dimension. The effective electric field is defined to produce the same energy transfer as a dc field and is given by

$$E_e = \frac{E\nu_c}{\sqrt{\nu_c^2 + \omega^2}} \tag{2}$$

where

E is the rms electric field (volts/cm)
ν_c is the collision frequency (collisions/sec)
ω is the radian frequency of the applied field (radians/sec)

If the near-zone field region about an antenna does vary with varying characteristic dimension then E_e/p is no longer simply a function of pressure times the characteristic dimension. This occurs, for example, with slot antennas when the slot width becomes appreciable compared to a wavelength.

When diffusion losses are negligible, CW breakdown occurs when the ionization rate is equal to the attachment rate. Since neither of these quantities depends upon geometry, but only on the values of field strength and pressure, the breakdown field will be independent of geometry. In pulsed breakdown, when the pulse is on

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† Stanford Res. Inst., Menlo Park, Calif.

[1] J. Chown, W. Scharfman and T. Morita, "Voltage breakdown characteristics of microwave antennas," Proc. IRE, vol. 47, no. 8, pp. 1331–1337; August, 1959.

[2] S. C. Brown, "High frequency gas discharge-breakdown," Proc. IRE, vol. 39, pp. 1493–1501; December, 1951.

[3] L. Gould and L. W. Roberts, "Breakdown of air at microwave frequencies," *J. Appl. Phys.*, vol. 27, pp. 1162–1170; October, 1956.

[4] H. Margenau, "Theory of high frequency gas discharges. IV. Note on the similarity principle," *Phys. Rev.*, vol. 73, no. 4, pp. 326; February 15, 1948.

for such a short time that the diffusion losses are small, the breakdown field tends to be independent of geometry. In this case, breakdown will occur when the difference between the ionization rate and attachment rate is large enough so that within the time that the pulse is on, the electron density will build up to the breakdown density. When diffusion losses are not negligible, the geometry will play an important part in determining the breakdown level.

The pressure below which diffusion losses become important will vary with antenna structure but, in general, at high pressures or when the characteristic dimension of the structure (such as the spacing in parallel plate transmission line) is large, the main loss mechanism will be attachment. When the pressure is low, or the characteristic dimension is small, the main loss mechanism will be diffusion. Thus, in typical plots of breakdown field as a function of pressure, the high pressure region is usually identified with attachment-controlled breakdown, while the low pressure region is identified with diffusion-controlled breakdown. However, at pressures so low that the mean free path is of the order of, or greater than the characteristic dimension, the process is no longer diffusion-controlled, and (1) no longer applies.

For breakdown in air the diffusion coefficient D_F is due to diffusion of free electrons and is given by

$$D_F = \mu_e \frac{kT_{ef}}{e} \tag{3}$$

where

μ_e = electron mobility $= e/m\sqrt{\nu_c^2+\omega^2}$
k = Boltzmann's constant
T_{ef} = electronic temperature with free diffusion.

A phenomenon which is unique to high altitude missile operation is the transmission of signal through the ionized media associated with reentry into the earth's atmosphere. The effect of the ionized media is to change the electron loss mechanism from free diffusion to ambipolar diffusion.[5] The diffusion of electrons from a neutral plasma brings about the increased mobility of the heavy positive ions, since the plasma tends toward charge neutrality. A positive space charge is developed as a sheath around the antenna, which retards the high velocity electrons and accelerates the positive ions. In the plasma to be considered in this paper, the electron temperature is large compared to the positive ion temperature, and the ambipolar diffusion coefficient is then given by

$$D_A \approx \mu_+ \frac{kT_{ea}}{e}, \tag{4}$$

where

μ_+ = positive ion mobility
T_{ea} = electronic temperature with ambipolar diffusion.

Thus

$$\frac{D_A}{D_F} = \frac{\mu_+ T_{ea}}{\mu_e T_{ef}}. \tag{5}$$

The net result will be to decrease the diffusion rate of electrons, thus decreasing the power-handling capability in an ionized medium.

III. Experimental Results

A 4-foot-diameter Plexiglass sphere was used for the experiments. This chamber can be evacuated to 4-micron Hg pressure which corresponds to an altitude of about 300,000 feet. The incident, reflected, and transmitted power are monitored as well as the pulse shape, frequency spectrum, and radiation pattern.[1]

A. Monopole Antenna

The effect of antenna height and diameter on breakdown power of a monopole over a conducting ground plane has been investigated with CW power at 240 and 400 mc, and with pulse power at 9400 mc. Fig. 1 shows the variation of minimum breakdown power of a monopole antenna as a function of antenna height at a fixed frequency of 240 mc and a pressure of 0.3 mm Hg. Measurements were made on antennas that were 0.08, 0.16, 0.24, 0.32, 0.40 wavelength long. As would be expected, the shortest antennas required the least power to produce breakdown, since they produce the greatest field strength per unit input power. All of the antennas broke down near the antenna tip, except the 0.40-wavelength monopole, which broke down at the feed point. This is reasonable from a consideration of the distribution of charge along the monopoles. For all but the 0.40 λ antenna, the maximum charge is at the tip. For the 0.40 λ antenna, it is at the feed point. Compu-

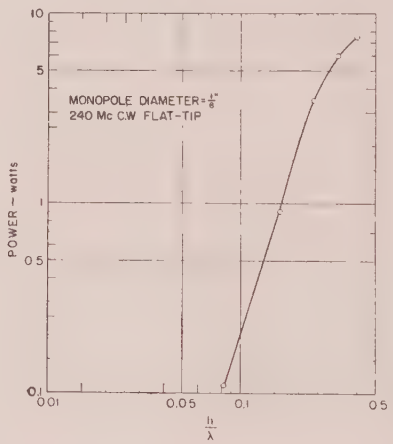

Fig. 1—Power to initiate breakdown as a function of antenna height at minimum pressure (approximately 0.3 mm Hg).

[5] W. P. Allis and D. J. Rose, "The transition from free to ambipolar diffusion," *Phys. Rev.*, vol. 93, no. 1, pp. 84–93; January 1, 1954.

tations made of the field at the monopole tip, for those antennas that broke down at the tip, shows that the breakdown fields were identical within the accuracy of the computation and measurements.

The fact that the fields required for breakdown were the same for these antennas, while the powers were very different, illustrates the fallacy of considering only the electric field in designing antennas to have large power-handling capabilities. This will be discussed later.

The effect of antenna diameter on breakdown power is illustrated in Fig. 2. Breakdown power, as a function

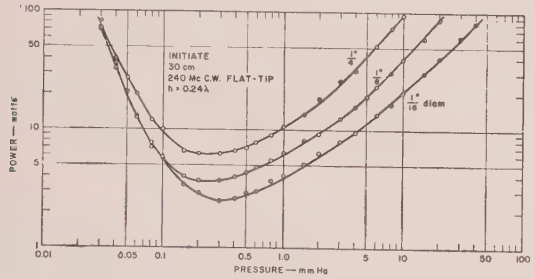

Fig. 2—Power to initiate breakdown as a function of pressure for different monopole diameters.

of pressure for monopole diameters of $\frac{1}{16}$, $\frac{1}{8}$, and $\frac{1}{4}$ inch, was measured for a 0.24 λ monopole. At higher pressures, the larger-diameter monopoles required more power than the smaller-diameter monopoles to produce breakdown. As the pressure was lowered, the curves merged, although the smallest diameter monopole required the lowest amount of power to produce breakdown at the pressure minimum. This behavior may be understood qualitatively by considering the electric field distribution for the various diameters. Assuming the impedance change with diameter is negligible, the total charge at the monopole tip for a one-watt input at the feed terminals will be approximately the same for monopoles of different diameters. However, the maximum electric field at the monopole surface will be inversely proportional to the diameter. Therefore, the thickest monopole will have the lowest surface electric field per watt of input. At high pressures, the radial extent of the region in which ionization occurs is small compared to the radius of the monopole, so that the field is relatively constant throughout the region. For this condition, the important parameter is the value of the electric field at the surface of the antenna. Therefore, the monopole of large radius requires more power for breakdown than the thin monopole. At low pressures, both because of increased electron diffusion and reduced electron production per unit volume, the radial dimension of the volume in which ionization occurs becomes substantially greater than the radius of the monopole. For this situation, the radius of the antenna is of small importance, and monopoles of different radii tend to require the same power for breakdown.

When the measured data on monopoles of different heights and diameters, and at different frequencies, are normalized and plotted as shown in Fig. 3, it becomes clear that scaling laws are applicable to monopole antennas over a wide range of parameters. The curve in Fig. 3 is the average for all the measured data. The deviation from the average is no greater than 20 per cent for values of pr_1 (r_1 is the monopole radius in cm) down to 0.05 and no greater than 30 per cent down to pr_1 equal to 0.005. Measurements were not made at sufficiently high values of pr_1 so that the value of E_e/p was independent of pr_1, although such a point is being approached. At that point, the breakdown would be attachment-controlled. For the highest values of pr_1 shown, the breakdown is controlled by a combination of attachment and diffusion, while at lower pr_1 diffusion is the controlling loss mechanism.

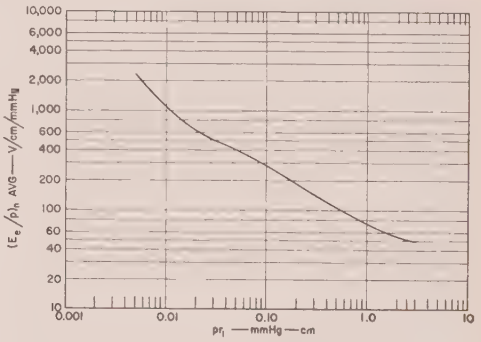

Fig. 3—Average of normalized monopole data.

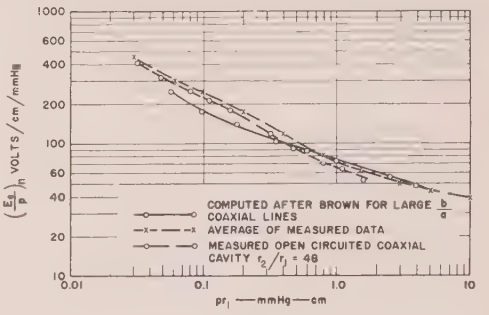
$(\frac{E_b}{p})_h$ VOLTS/cm/mm Hg
COMPUTED AFTER BROWN FOR LARGE $\frac{b}{a}$
COAXIAL LINES
AVERAGE OF MEASURED DATA
MEASURED OPEN CIRCUITED COAXIAL CAVITY $r_2/r_1 = 48$
pr_1 — mm Hg — cm

Fig. 4—Coaxial line and monopole breakdown data.

For purposes of breakdown calculation, the concept of a monopole as a coaxial line with large ratio of outer to inner conductor radius was applied to obtain an approximate solution for (1), the electron continuity equation.[6] The results of the calculation are plotted in Fig. 4. Also plotted for comparison is the curve of the average of the measured monopole data, as well as the measured data for an open-circuited coaxial line with $b/a = 48$. The agreement is reasonably good, considering the uncertainty of the measured data and the approximation required for the analytic solution. For values of

[6] M. A. Herlin and S. C. Brown, "Electrical breakdown of a gas between coaxial cylinders," *Phys. Rev.*, vol. 74, no. 8, pp. 910–913; October 15, 1948.

E_e/p much above 100, the approximations used in the analytical solution become increasingly inaccurate.

The behavior of a monopole antenna subjected to high pulsed power was investigated at 9400 mc. Pulse widths of 0.5, 1.0 and 2.3 μsec were used. Over this range of pulse widths, there was no significant variation of the values of E_e/p, when plotted as a function of $p\tau$ for each value of pr_1. In fact, the measured values of E_e/p for pulsed power were very close to the values measured at 240 and 400 mc with CW power.

Fig. 5 is the plot of the power required to initiate and maintain breakdown on a λ/4 monopole. For CW measurement, the power required to initiate breakdown is about 6 to 9 db higher than the power required to maintain breakdown. For initiation of breakdown, free diffusion is the dominant loss mechanism, while for maintaining breakdown, ambipolar diffusion is the dominant loss mechanism.

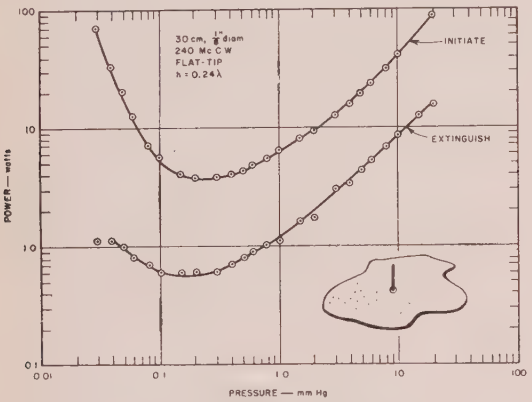

Fig. 5—Power to initiate and extinguish breakdown as a function of pressure.

B. Aperture Antennas

1) CW breakdown. Measurements made on a narrow slot antenna with a dielectric cover at 380 mc show that the electric fields required to initiate breakdown are considerably higher than for a rectangular waveguide of the same cross section. This is shown in Fig. 6. The loss of electrons to the dielectric cover is much greater than the loss to the walls in the waveguide case. Furthermore, the fast fall-off of the near-zone field, measured normal to the aperture, also increases the diffusion loss.

Also shown in Fig. 6 is the field strength necessary to maintain breakdown. This curve was calculated by assuming that the field distribution near the aperture has not changed from the no breakdown condition to the maintaining condition.

2) Pulsed breakdown. Pulsed breakdown of slot antennas was investigated at 9400 mc. A series of measurements was made of the power required to initiate breakdown, as a function of pulse width and repetition rate, as the pressure was varied. The results of the

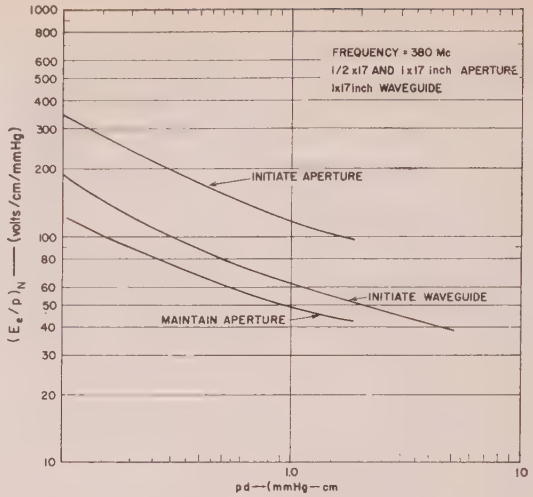

Fig. 6—CW aperture and waveguide breakdown.

repetition rate studies showed that for the usual rates in radar systems, breakdown is a single pulse phenomenon, extinguishing at the end of each pulse. The repetition rate had only a small effect on the power required to initiate power for rates up to 1000 pps. The effect of slot height is shown in Fig. 7, where the power to initiate breakdown is plotted as a function of pressure for slot heights of 0.050, 0.100, and 0.400 inch. In all cases, the aperture was covered with a dielectric cover. Fig. 7 is for approximately 0.5-μsec pulse width. Similar measurements were made with 1.0- and 2.3-μsec pulse widths. These data have been normalized and are shown plotted in Fig. 8. The normalized data should be useful in estimating breakdown fields, for a wide range of parameters for narrow slot antennas. The dashed curves for the 0.4-inch aperture are beginning to deviate from the narrower slot data, indicating that the near zone field distribution is changing with b.

Fig. 9 is a plot of E_e/p plotted vs $p\tau$, where τ is the pulse width, for an X-band waveguide radiator, with dielectric cover and for an X-band waveguide. Here the parameter, pb, is the pressure times the guide height. For large pb the two curves converge since attachment is the principal loss-mechanism. Furthermore they converge toward the CW value which is shown on the right.[2,3,7] For small pb, the diffusion to the cover greatly increases the breakdown field for the slot radiator with the dielectric cover.

C. Antenna Operation in Weakly Ionized Media

The effect on the breakdown power of an antenna, when its environment includes a plasma, has been investigated. A dc glow discharge was formed with the negative glow area of the discharge located over the

[7] F. Worth, "A Study of Voltage Breakdown in the Cavity Fed Slot Antenna," Missile Systems Div., Lockheed Aircraft Corp., Sunnyvale, Calif., Rept. MSD 2030; January, 1957.

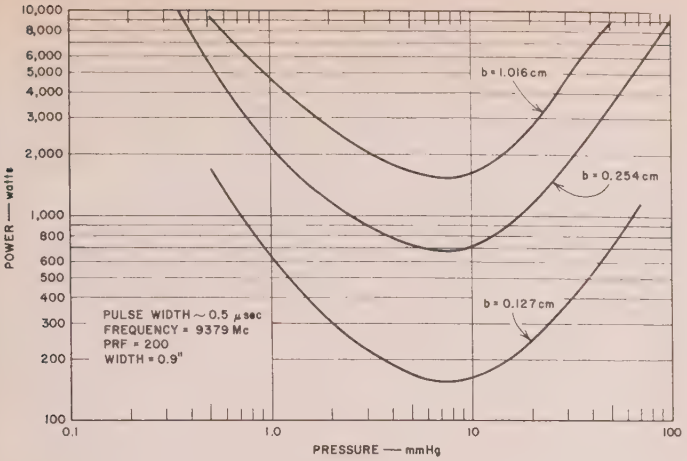

Fig. 7—Pulse power to initiate breakdown as a function of pressure for different height apertures.

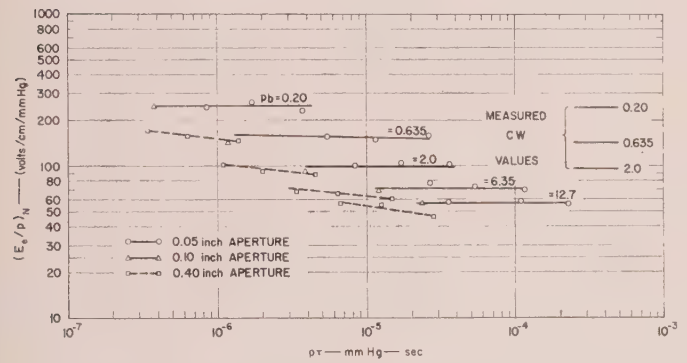

Fig. 8—Normalized pulse data for narrow slot antenna from measured data at 9379 mc.

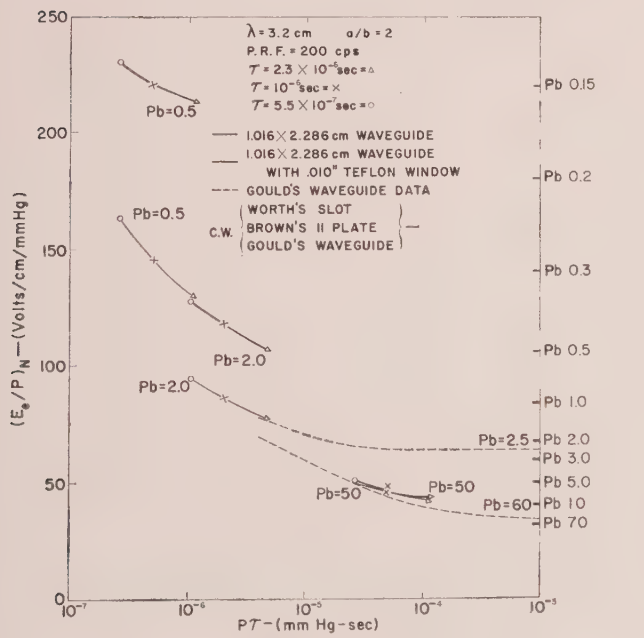

Fig. 9—Comparison of X-band pulsed breakdown for waveguide and for an aperture of the same dimensions.

aperture of a slot antenna radiating at 373 mc. While the density of electrons in this plasma is of the same order of magnitude as for a reentry body, the gas temperature itself is very low. In the negative glow area, the plasma is essentially neutral with a maximum electron density of 10^{11} electrons per cc decreasing to zero over the antenna aperture.[8] Figs. 10 and 11 show the results obtained when the effective plasma frequency ($f_p = 8900\sqrt{N}$) is approximately one-half and equal to the RF frequency. The plasma frequency was obtained from the transmission and reflection measurement prior to RF breakdown, with the collision frequency assumed equal to the RF frequency, since the pressure was chosen so that the RF radian frequency was approximately equal to the collision frequency before breakdown.

For Fig. 10, where the plasma frequency is approximately one-half the RF frequency, the attenuation is negligible and the transmitted and reflected power increase directly with the incident power. After the occurrence of voltage breakdown at 4.5 watts, the transmitted power remains almost constant, while the absorbed power increases directly with the incident power. The transition to RF breakdown is shown to be double-valued, since the conditions abruptly change at breakdown. For Fig. 11, where the plasma frequency is approximately equal to the RF frequency, the characteristic after RF breakdown is almost identical to that given in Fig. 10. However, before RF breakdown occurs, there is RF attenuation and the VSWR is higher than for Fig. 10. Before RF breakdown, the RF conductivity appears to be constant, independent of the applied field. In the absence of a plasma, this antenna required about 36 watts for breakdown as compared to 7 watts with a plasma.

Similar measurements have been made with pulsed power at X-band with a dc discharge, and with a 300-mc RF discharge used as a source of ambient ionization. An X-band waveguide 0.4 by 0.9 inch, with a dielectric window, was used as the antenna. The electron density produced by either discharge was sufficiently low, that negligible absorption of the X-band signal was observed before X-band breakdown. The results with the 300-mc discharge are shown in Fig. 12 for a pressure of 5 mm Hg. X-band breakdown occurred at a power level of 1.0 kw and was evidenced by the removal of a small part of the end of the transmitted pulse. Further increases in input power resulted in less and less of the incident pulse being transmitted.[1] The portion of the input pulse that did not appear in the transmitted signal was absorbed by the plasma and is plotted as the absorbed power. Note that the absorbed power curve is

[8] J. D. Cobine, "Gaseous Conductors, Theory and Engineering Applications," Dover Publications, Inc., New York, N. Y., pp. 212–226; 1958.

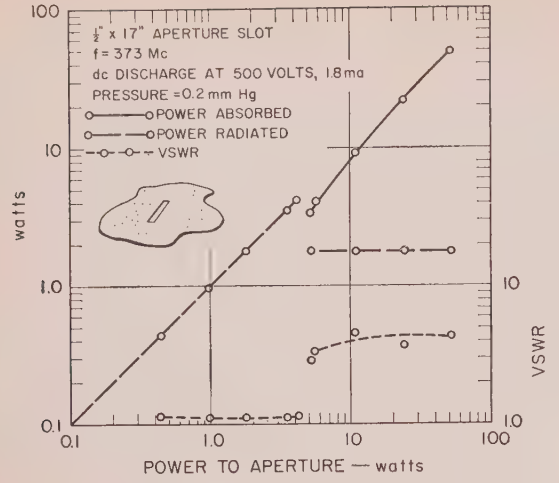

Fig. 10—Power handling characteristics of slot antenna in the presence of a 500 volt dc discharge.

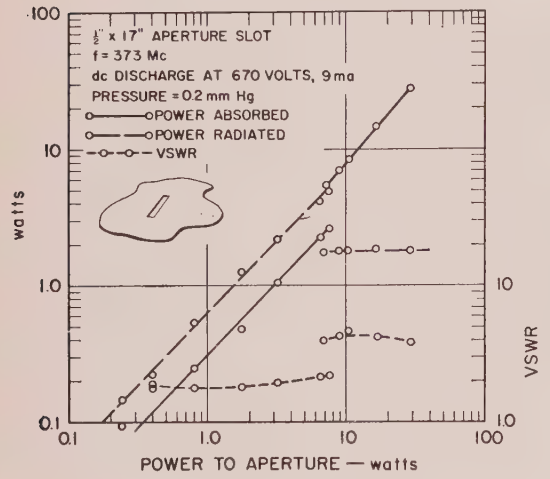

Fig. 11—Power handling characteristics of slot antenna in the presence of a 670 volt dc discharge.

approaching a line of slope equal to unity. That is, conditions are approaching the case in which all of the input power is being absorbed. Before X-band breakdown, absorption would have been observed as a decrease in the level of the transmitted pulse when the 300-mc discharge was turned on. Within an accuracy of about 0.1 db, no decrease in level was observed. Thus the absorption before X-band breakdown was negligible.

Also plotted in Fig. 12 is the VSWR as a function of input power. Again, there was no change in VSWR before X-band breakdown, but after X-band breakdown the VSWR rises with input power. The VSWR shown in this figure is found from a value of reflection coefficient, which is calculated by measuring the ratio of the power in the reflected pulse, averaged over one pulse width, to the power in the incident pulse, averaged over one pulse width. Thus, in a sense, this is an average VSWR. The instantaneous VSWR, considered as calculable from the ratio of the instantaneous value of the reflected pulse to the instantaneous value of the in

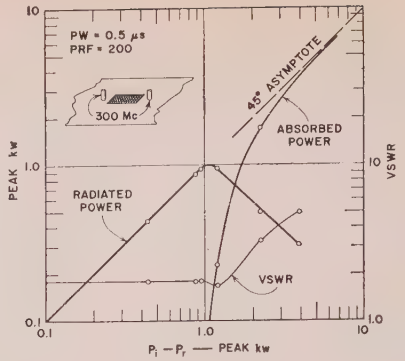

Fig. 12—X-band aperture with 300 mc discharge; $p_0 = 5.0$ mm. Hg.

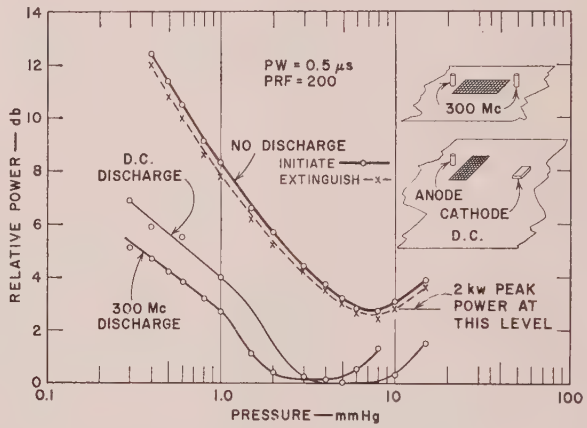

Fig. 13—Power to initiate X-band aperture for different conditions of ambient ionization.

cident pulse, may be considerably higher than the average shown in Fig. 12. Which value of VSWR is significant will depend upon the particular system for which the antenna is designed.

The effect of the 300-mc discharge upon the breakdown level of the X-band antenna was measured as a function of pressure. The results are shown in Fig. 13. Since the extent of the 300-mc discharge varied with pressure, the results are included only to indicate that a plasma may severely influence the breakdown level of pulsed transmission. A much more refined experiment and theory of breakdown would be required to relate the plasma properties to the onset of breakdown.

One interesting result of these studies is that the CW power-handling capability of the slot, in the presence of an ionized medium, appears to be about the same as the power required to maintain the breakdown. Since the main loss mechanism is the same for both cases, this would be true if the RF field distribution is approximately the same and the energy of the electrons produced by the dc discharge did not affect the RF breakdown. It is clear that the diffusion process changes from free to ambipolar in the presence of strong ambient ionization. However, more work is needed to separate the effects of the energy of the ambient ionization from the effects of the density itself. The results of the measurements of the pulse power-handling capabil-

ity of a slot antenna, in the presence of an ionized plasma, showed that the initiating levels were reduced appreciably below the maintaining level in the absence of a plasma. This is quite different than for the CW measurements. For the pulse case, the difference between the initiating and the maintaining level is due to the residue of electrons left from the previous broken down pulse. If the conditions are such that the electron density from the previous broken down pulse is negligible, then there will be a negligible difference between the initiating and maintaining level; both will be operating with a free diffusion condition. The fact that the initiating and maintaining levels are so close to each other indicates that a low electron density is present at the beginning of the maintaining pulse. With a plasma present, a high electron density is present at all times. The electron density, at the beginning of the maintaining pulse, is much higher than when no plasma is present; hence the diffusion loss will tend to be ambipolar. The electric field required to maintain breakdown is therefore lower.

D. Methods for Increasing the Power-Handling Capability of Antennas

There are several methods by which the power-handling capability of an antenna may be increased. One method is to increase the electron-loss mechanism. This may be accomplished by the use of a dc bias, whether negative or positive, to attract the electrons to the antenna or repel the electrons from its vicinity. Fig. 14 shows the large increase in power-handling capability obtained for a $\lambda/4$ monopole with a dc bias.[9] These results were obtained by applying a dc voltage between the ground plane and the monopole. The breakdown power was measured for different values of dc voltage. Negative voltages indicate that the monopole was negative with respect to the ground plane. Similar results are obtained for a narrow slot radiator. However, as the slot width is increased, the near-zone field drop-off decreases considerably, and the dc field is less effective. For example, for an *X*-band waveguide aperture with a dielectric cover, only 3 db increase in power handling capability could be obtained with various biasing arrangements at the pressure for minimum power-handling capability. It should be pointed out that dc bias will not be effective with an antenna imbedded in a dielectric. A charge will build up on the dielectric so that the dc voltage will appear across the dielectric, rather than in the region where electrons are being formed.

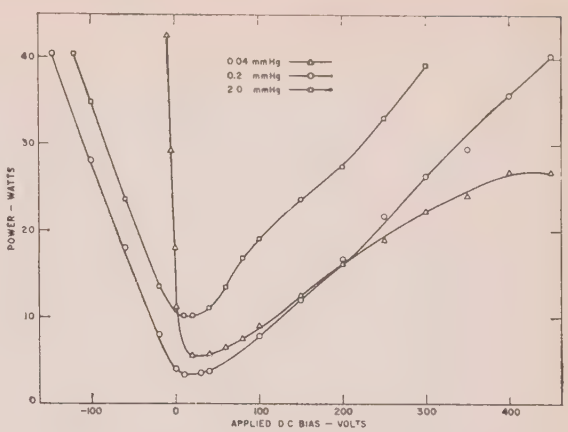

Fig. 14—Power to initiate voltage breakdown as a function of applied dc biasing voltage for a 0.25 wavelength monopole.

Another method for increasing the power-handling capability is to decrease the electric field strength and thus decrease the ionization rate. The use of a pressurized radome over the antenna structure to a point where the field has decreased sufficiently is a possibility. Another possibility is the proper choice of antenna configuration. A top-loaded monopole, only 0.10 λ in height, has been designed and constructed which handles 8.5 watts as compared to 0.20 watt for a $\frac{1}{8}$-inch-diameter monopole of the same height.

Increasing the *field strength* required for breakdown may not always be desirable, since it is the maximum *radiated* power which is desired and not maximum near-zone field strength. As an illustration, consider the case of an open waveguide slot with no dielectric cover. The field strength for breakdown can be increased by decreasing the slot width, since the loss of electrons to the wall is increased as the walls are brought closer together. However, the increase in field strength per watt is faster than the increase in breakdown field strength, so that the power handling capability decreases even though the field strength for breakdown is increased.

The factors that affect the design of antennas that are to have high power-handling capabilities in the absence of a plasma have been discussed in this paper. Further work is needed to determine antenna design procedures for antennas to have high power-handling capabilities in the presence of a plasma.

IV. Acknowledgment

The authors gratefully acknowledge the assistant of W. Wadsworth in the performance of the experimental work.

[9] E. Blasi, Lockheed Missiles and Space Div., Sunnyvale, Calif., has reported to the authors that this technique has been applied with successful flight results to an antenna for use on a ballistic missile.

A Small-RF-Signal Theory for an Electrostatically Focused Traveling-Wave Tube*

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Summary—**This paper describes a small-RF-signal theory for a new type of electrostatically focused traveling-wave tube. Consideration is given to the effect on the gain of large periodic variations of dc electron velocity, resulting from electrostatic focusing. Llewellyn-Peterson equations which apply to dc voltage distributions, similar to those used in the electrostatically focused traveling-wave tube, are used for analysis. The limitation of the Llewellyn-Peterson equations to one-dimensional problems is overcome by use of correction factors derived from Pierce's theory for traveling-wave tubes. Generalized expressions taking into account the periodic variations of dc electron velocity are derived for the gain parameter C and space-charge parameter QC. For uniform dc electron velocity, such as that found in a magnetically focused tube, these expressions reduce to the forms derived by Pierce. In addition to their suitability for analysis of electrostatically focused traveling-wave tubes, the Llewellyn-Peterson equations provide a method of analyzing magnetically focused traveling-wave tubes alternative to those used previously. It is shown that the gain of the electrostatically focused tube can be calculated by use of the gain and space-charge parameters of the electrostatically focused tube and graphical data available for magnetically focused tubes. Depending upon the voltage distribution in the region of electron flow, the gain and bandwidth of a traveling-wave RF circuit employing an electron beam having a periodically varying dc velocity can be greater or smaller than the gain and bandwidth of the same RF circuit employing a uniform dc velocity electron beam. Measurements performed on an experimental electrostatically focused tube are in agreement with the theory.**

Introduction

THIS paper presents a theory of small-signal RF interaction between an electron beam having a periodically varying dc velocity, and a traveling-wave type circuit. The conditions analyzed approximate those existing in a new type of experimental electrostatically focused traveling-wave tube. Development of such a tube was undertaken, primarily, to produce a tube package considerably lighter than that of a tube having comparable RF characteristics, but employing magnetic fields for electron beam focusing. Several authors have described electrostatically focused traveling-wave tubes which use bifilar helices as the RF circuit and beam-focusing elements [1]–[5]. Application of helices is limited, however, to relatively low peak and average RF power levels [6], [7]. Fig. 1 shows schematic representations of two RF circuits suitable for electrostatic beam focusing, and for high peak-RF-power applications [8]. These structures, which are basically the "folded-line" and "capacitively-coupled-bar" circuits [9], [10], are modified for electrostatic beam focusing by the use of focusing electrodes at the center of the interaction regions, as shown in Fig. 1. These focusing electrodes are maintained at a dc voltage V_b, and are insulated from the main body of the RF circuit and from the cylindrical drift tubes, which are supplied with a dc voltage V_a. Consequently, they establish a spatially varying dc potential to provide the desired focusing action in the region of electron flow [11].

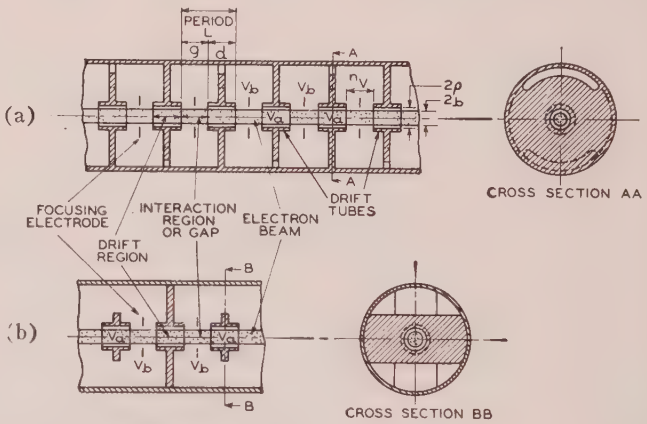

Fig. 1—Schematic representations of two high-power traveling-wave tube circuits modified for electrostatic beam focusing: (a) "folded line" circuit, (b) "capacitively-coupled bar" circuit.

Each focusing electrode is supported by an RF choke mounted on ceramic insulators which are brazed to the body of the RF circuit. The RF chokes prevent radiation of RF energy through the ceramic insulators across the range of operating frequencies [8].

A study performed by Belohoubek [8] has shown that the electrode configuration shown in Fig. 1 can be incorporated most easily in a space-harmonic structure [9]. This structure is well suited for electrostatic focusing because the length g of the interaction region is sufficiently large to prevent dc voltage breakdown between the focusing electrodes and the drift tubes.

Because forward-wave amplification in space-harmonic structures is most efficient when the electron dc transit angle per period is in the range from π to 2π radians [9], the graphical results presented in this paper are plotted for this range only. However, the theory and mathematical analysis apply to any value of dc transit angle per period.

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Assumptions and Method of Analysis

For the purpose of this analysis, it is assumed that the actual RF structures may be represented by an equivalent circuit consisting of a periodically loaded smooth transmission line, as shown in Fig. 2(a).

The following is a brief description of the idealized properties of the RF circuit prior to the introduction of the electron beam. For simplicity, only the axial RF electric field is considered in the region of electron flow. It is assumed that this field is independent of the spatial coordinates in each gap and varies in the nth gap as $E_n e^{j\omega t}$, where E_n denotes the complex amplitude of the field in the nth gap, t is the time variable, and $\omega = 2\pi \times$frequency. If an RF generator is connected to the left of the nth interaction region, the RF electric field in the $(n+p)$th interaction region is $E_n e^{j(\omega t - p\theta)}$, where θ is the phase shift per period in the absence of the electron beam. It is further assumed, that in the absence of space charge, the RF fields in the drift regions are zero. The idealized RF electric field distribution for an infinitely thin focusing electrode, real E_n, and for $\theta = 3\pi/2$ is shown in Fig. 2(b).

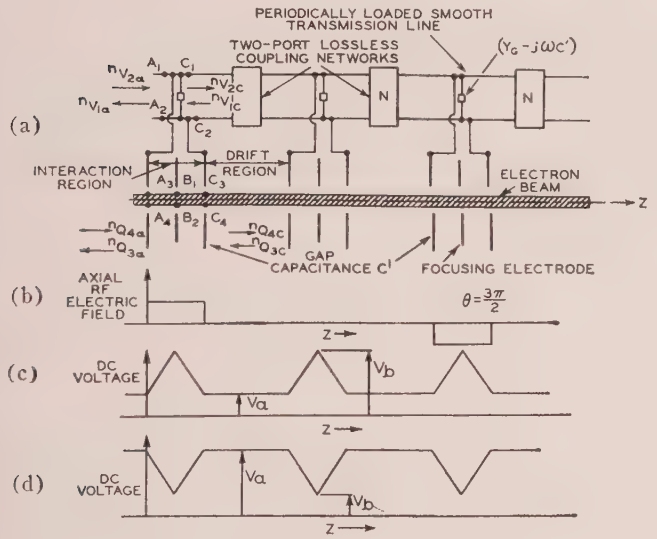

Fig. 2—(a) Assumed equivalent circuit of the electrostatically focused high-power traveling-wave tube. (b) Idealized, one-dimensional, axial, RF electric-field distribution for $\theta = 3\pi/2$, prior to the introduction of the electron beam. (c) and (d) Idealized, one-dimensional dc-voltage distributions in the absence of the electron beam.

The idealized RF electric-field distribution described above is that which exists in a gridded structure having a negligible variation of RF fields in the transverse direction. A correction factor which takes into account the variation of RF fields in the transverse direction will be introduced later, to give an approximate analysis for gridless structures. The one-dimensional dc voltage distributions shown in Figs. 2(c) and 2(d) are simplified representations of the voltage distributions which can be used for electrostatic beam focusing.

It is assumed that the transverse dimensions of the electron beam are constant, as shown in the lower portion of Fig. 2(a). In a well-designed electrostatically focused traveling-wave tube, the perturbations in the transverse dimenions of the electron beam are small. Theoretically, it is possible to design focusing electrodes capable of maintaining constant transverse beam dimensions [12].

The theory of RF interaction between the RF circuit and the electron beam differs in this case from the usual theories [13]–[15], because the dc electron velocity has a large, periodic, spatial variation resulting from electrostatic beam focusing. Llewellyn-Peterson equations [16], [17], which apply to the dc voltage distributions shown in Fig. 2(c) and 2(d), are used in this paper to analyze the RF interaction between the electron beam and the RF circuit. The Llewellyn-Peterson equations apply to a one-dimensional model of electron flow normal to planes or grids which can be maintained at different dc potentials. Electrostatically focused traveling-wave tubes do not usually employ gridded gaps, but the Llewellyn-Peterson equations can be combined with space-charge [18] wave theory and Pierce's [13], [14] theory for traveling-wave tubes to give an approximate analysis for gridless structures.

The initial phase of the analysis is the determination of the RF beam modulation and RF gap voltages at the input of a typical interaction region in terms of the corresponding quantities at the input of the preceding interaction region. This procedure leads to four linear, simultaneous, finite difference equations with constant coefficients [19] from which the small-signal gain per unit cell and, hence, the over-all gain of the tube can be calculated.

In this paper, mks units are used.

Approximate Analysis of RF Gap Interaction

The first phase of this analysis is the determination of the RF quantities at the output of the nth interaction region in terms of the corresponding quantities at the input of the same region.

It is assumed that the RF gap voltage is the sum of the forward and reflected modes propagating on the smooth sections of the transmission line. The input and output reference terminals for these modes are marked in Fig. 2(a) by A_1A_2 and C_1C_2, respectively, and are located immediately to the left and right of the total gap admittance $(Y_G + Y_B)$, where Y_G is the gap admittance in the absence of the electron beam and Y_B is the beam loading admittance. The beam loading admittance Y_B is in parallel with the capacitance C' shown in Fig. 2(a). The symbol C' denotes the capacitance of the interaction region in the vicinity of the electron beam. Because the present analysis is limited to relatively low space-charge effects, the beam loading admittance Y_B does not appear in the final results, and mathematical expressions for that quantity will not be given.

It is convenient to perform the analysis in terms of the normalized RF quantities ${}^nQ_1$ and ${}^nQ_2$ used by Heffner [20] and defined as follows:

$$ {}^nQ_{1a,c} = \frac{{}^nV_{1a,c}}{\sqrt{2Z_0}} \tag{1} $$

$$ {}^nQ_{2a,c} = \frac{{}^nV_{2a,c}}{\sqrt{2Z_0}}, \tag{2} $$

where Z_0 is the characteristic impedance of the smooth sections of the transmission line. The subscripts a and c denote, respectively, reference to terminals A_1A_2 and C_1C_2 shown in Fig. 2(a) . The subscripts 1 and 2 denote, respectively, the peak values of reverse and forward components of the RF gap voltage measured across the gap at radius ρ as shown in Fig. 1(a).

The input and output reference terminals for the RF beam modulation are, respectively, A_3A_4 and C_3C_4, and coincide with the input and output of the interaction region shown in Fig. 2(a). The RF modulation on the electron beam will be described in terms of normalized space-charge wave quantities [21, 22] ${}^nQ_3$ and ${}^nQ_4$, defined as follows:

$$ {}^nQ_{3a,c} = \sqrt{\frac{V_a\omega_a}{I_0\omega}}\, i_{sa,c} \tag{3} $$

$$ {}^nQ_{4a,c} = \sqrt{\frac{V_a\omega_a}{I_0\omega}}\, i_{fa,c} \tag{4} $$

where i_s and i_f denote, respectively, the peak values of the RF convection current in the slow and fast space-charge waves, V_a is the dc voltage in the drift region, I_0 is the dc beam current, and ω_a is the angular plasma frequency [23] given by:

$$ \omega_a = \sqrt{\frac{\eta I_D}{\epsilon u_a}}\,. \tag{5} $$

In this expression, I_D is the dc beam current density, η is the charge-to-mass ratio of the electron, ϵ is the dielectric constant of free space and

$$ u_a = \sqrt{2\eta V_a}. \tag{6} $$

The superscript n in (1)–(4) denotes reference to the nth interaction region. The following relationships result when Heffner's [15] (1), (2), (7), and (10) are combined:

$$ q_{a,c} = \frac{1}{A}\,(i_{sa,c} + i_{fa,c}) \tag{7} $$

$$ v_{a,c} = \frac{u_a\omega_a}{I_0\omega}\,(i_{fa,c} - i_{sa,c}), \tag{8} $$

where q denotes the peak value of the total RF convection current density, v is the peak value of RF electron velocity, and A is the cross-sectional area of the electron beam. The subscripts a and c in (1)–(4), (7), and (8) must be taken in the order indicated. In terms of normalized quantities, the expression for RF input power P_B carried by the electron beam, assumes the following simple form [24], [25]:

$$ P_B = {}^nQ_{4a}{}^nQ_{4a}{}^* - {}^nQ_{3a}{}^nQ_{3a}{}^*. \tag{9} $$

The first phase of the analysis is the determination of the elements of the inverse transfer [26], [27] matrix (a) defined as follows:

$$ \begin{pmatrix} {}^nQ_{1c} \\ {}^nQ_{2c} \\ {}^nQ_{3c} \\ {}^nQ_{4c} \end{pmatrix} = (a) \begin{pmatrix} {}^nQ_{1a} \\ {}^nQ_{2a} \\ {}^nQ_{3a} \\ {}^nQ_{4a} \end{pmatrix}. \tag{10} $$

When the RF gap voltage is zero

$$ ({}^nQ_{1a} + {}^nQ_{2a} = {}^nQ_{1c} + {}^nQ_{2c} = 0), $$

and the radius b of the electron beam which moves at a uniform dc velocity u_0 is large (ideally infinite), the output RF beam modulation is given by the following relationships [18, 28]:

$$ {}^nQ_{3c} = {}^nQ_{3a}e^{-j\omega_p g/u_0}e^{-j\omega g/u_0} \tag{11} $$

$$ {}^nQ_{4c} = {}^nQ_{4a}e^{j\omega_p g/u_0}e^{-j\omega g/u_0}, \tag{12} $$

where g is the length of the interaction region and ω_p is the value of ω_a, given by (5), when u_a is equal to u_0.

Let parameters β_a and β_{pa}, associated with the dc velocity in the drift regions of the electrostatically focused tube, be defined by the following relationships:

$$ \beta_a = \frac{\omega}{u_a} \tag{13} $$

and

$$ \beta_{pa} = \frac{\omega_a}{u_a}\,. \tag{14} $$

It is shown in Appendix I that when the dc voltage distribution is as shown in Fig. 2(d) or 2(c), and the total RF current density [29] I is zero, the relationships between the output and input normalized space-charge wave quantities are as follows:

$$ {}^nQ_{3c} = a_{33}\,{}^nQ_{3a} + a_{34}\,{}^nQ_{4a} \tag{15} $$

$$ {}^nQ_{4c} = a_{43}\,{}^nQ_{3a} + a_{44}\,{}^nQ_{4a}. \tag{16} $$

The coefficients a_{33}–a_{44} in (15) and (16) are given by (103)–(105) in Appendix I for small values of the space-charge factor [30] ζ. The variation of RF beam modulation along an electron beam moving with uniform dc velocity through a conducting cylinder can be analyzed, approximately, by means of the Llewellyn-Peterson equations. It has been shown [31], [32] from the Hahn [33] and Ramo [18] theories that if the transverse dimensions of such a beam are infinite, the total axial RF current density I is zero. Approximate analysis of RF modulation along such a beam can be obtained from the Llewellyn-Peterson equations by making the total

RF-current density zero, the space-charge factor ζ small ($\zeta \ll 1$), the dc velocities u_a and u_b equal to u_0, and ω_a equal to ω_p in (103)–(105). Under these conditions, (15) (16), (101), and (103)–(107), yield the following expressions:

$$ {}^nQ_{3c} = \left(1 - \frac{j\omega_p g}{u_0} - \frac{\omega_p^2 g^2}{u_0^2}\right) e^{-j\omega g/u_0}\, {}^nQ_{3a} \tag{17} $$

$$ {}^nQ_{4c} = \left(1 + \frac{j\omega_p g}{u_0} - \frac{\omega_p^2 g^2}{u_0^2}\right) e^{-j\omega g/u_0}\, {}^nQ_{4a}. \tag{18} $$

Comparison of (11), (12), (17) and (18) shows that the space-charge wave and Llewellyn-Peterson theories lead to almost identical results if $g \ll u_0/\omega_p$. The difference between the results obtained by the two theories is caused by the fact that the Llewellyn-Peterson equations take into account the voltage depression caused by space charge between the grids, whereas the dc voltage in a drift tube is assumed constant. The space-charge waves given by (11) and (12) can be derived from the Llewellyn-Peterson equations, as shown by Birdsall [34]. In a practical high-power traveling-wave tube, the length g of an interaction region is considerably smaller than the parameter u_0/ω_p, and application of (17) and (18) to such regions is, therefore, justified.

If the boundary of the perfectly conducting drift region of large transverse dimensions is interrupted periodically to form interaction gaps across which an RF circuit is connected, as shown in Fig. 1, a total RF current will, in general, circulate in the region of electron flow. If the initial RF beam modulation is zero (${}^nQ_{3a} = {}^nQ_{4a} = 0$), and the variation in the transverse direction is neglected, the total RF-current density I, in the region of electron flow, is as follows:

$$ I = \frac{{}^nV_{1a} + {}^nV_{2a}}{\overline{A}}, \tag{19} $$

where $\overline{A}$ is the impedance per unit area in the region of electron flow. Because the subsequent analysis is limited to small values of the space-charge factor ($\zeta \ll 1$), $\overline{A}$ is essentially equal to $g/j\omega\epsilon$.

It is shown in Appendix I that the combination of (19) with the Llewellyn-Peterson equations permits the RF beam modulation, caused by the RF gap voltage, to be expressed in terms of normalized RF quantities in the following manner:

$$ {}^nQ_{3c} = a_{31}\, {}^nQ_{1a} + a_{32}\, {}^nQ_{2a} \tag{20} $$

$$ {}^nQ_{4c} = a_{41}\, {}^nQ_{1a} + a_{42}\, {}^nQ_{2a}. \tag{21} $$

The coefficients a_{31}–a_{42} are given by (110) and (111) in Appendix I for small values of the space-charge factor ζ.

The total RF beam modulation at the output of the interaction region is the sum of components caused by the initial beam modulation given by (15) and (16), and RF-circuit voltages given by (20) and (21). In general, therefore,

$$ {}^nQ_{3c} = a_{31}\, {}^nQ_{1a} + a_{32}\, {}^nQ_{2a} + a_{33}\, {}^nQ_{3a} + a_{34}\, {}^nQ_{4a} \tag{22} $$

$$ {}^nQ_{4c} = a_{41}\, {}^nQ_{1a} + a_{42}\, {}^nQ_{2a} + a_{43}\, {}^nQ_{3a} + a_{44}\, {}^nQ_{4a}. \tag{23} $$

The elements a_{13}, a_{23}, a_{14}, a_{24} can be derived most easily from Heffner's relationships [35], which are obtained from the assumption of lossless exchange of RF power between the electron beam and the RF circuit. It is at this stage of the analysis that the use of normalized RF quantities facilitates evaluation of the desired coefficients. Fig. 2(a) shows that the RF gap input and output voltages are equal. Therefore,

$$ {}^nV_{1a} + {}^nV_{2a} = {}^nV_{1c} + {}^nV_{2c}. \tag{24} $$

Combination of (1), (2), (10), and (24) shows that the following relationships must be satisfied:

$$ a_{11} + a_{21} = 1 \tag{25} $$

$$ a_{12} + a_{22} = 1 \tag{26} $$

$$ a_{13} + a_{23} = 0 \tag{27} $$

$$ a_{14} + a_{24} = 0. \tag{28} $$

Eqs. (25) and (27) combined with Heffner's [15] (73) yield the expression

$$ a_{13} = a_{41}{}^* a_{43} - a_{31}{}^* a_{33}, \tag{29} $$

where * indicates the complex conjugate. If ω_a is sufficiently small so that terms involving $\omega_a{}^2$ are negligible compared to unity, combination of (29), (103), (104), (110) and (111) gives

$$ a_{13} = -a_{31} = K_0\left(1 - \omega_a\tau_i - j\frac{\omega_a g r}{2u_a}\right), \tag{30} $$

where r, and K_0 are given by (106) and (112), respectively. The parameter τ_i need not be defined because, as shown later, it does not appear in the final results.

Combination of (25) and (28) with Heffner's (74) produces the equation

$$ a_{14} = a_{41}{}^* a_{44} - a_{31}{}^* a_{34}. \tag{31} $$

Eqs. (31), (104), (105), (110) and (111), with the same assumptions as were made for (30), yield

$$ a_{14} = a_{41} = K_0\left(1 + \omega_a\tau_i + j\frac{\omega_a g r}{2u_a}\right). \tag{32} $$

The following equations can be derived by application of simple circuit theory [36].

$$ a_{12} = -a_{21}, \tag{33} $$

and

$$ a_{12} = \frac{Z_0}{2}(Y_G + Y_B). \tag{34} $$

An alternate but lengthier method of obtaining the elements in the first two rows of the (a) matrix is the use of Llewellyn-Peterson [16] (5); relating the RF gap voltage, total current density I and the beam modulation at the input of the interaction region.

Derivation of the Characteristic Equation for the Phase Shift Per Cell in Presence of the Electron Beam

The second phase of the analysis is the determination of the RF quantities at the input of the $(n+1)$th interaction region in terms of the corresponding quantities at the input of the nth interaction region. The quantities ${}^{n+1}Q_{1a}$ and ${}^{n+1}Q_{2a}$ are related linearly [37] to the quantities ${}^{n}Q_{1c}$ and ${}^{n}Q_{2c}$. Hence, in matrix notation,

$$\begin{pmatrix} {}^{n+1}Q_{1a} \\ {}^{n+1}Q_{2a} \end{pmatrix} = (b) \begin{pmatrix} {}^{n}Q_{1c} \\ {}^{n}Q_{2c} \end{pmatrix}. \tag{35}$$

The RF beam modulation is transformed through the drift region as follows [18], [28]:

$$\begin{pmatrix} {}^{n+1}Q_{1a} \\ {}^{n+1}Q_{4a} \end{pmatrix} = \begin{pmatrix} e^{-j\alpha} & 0 \\ 0 & e^{-j\mu} \end{pmatrix} \begin{pmatrix} {}^{n}Q_{3c} \\ {}^{n}Q_{4c} \end{pmatrix} \tag{36}$$

where

$$\alpha = (\beta_a + \beta_{pa})d \tag{37}$$

and

$$\mu = (\beta_a - \beta_{pa})d. \tag{38}$$

Eqs. (10), (35), and (36) can be combined to give the following relationship

$$\begin{pmatrix} {}^{n+1}Q_{1a} \\ {}^{n+1}Q_{2a} \\ {}^{n+1}Q_{3a} \\ {}^{n+1}Q_{4a} \end{pmatrix} = (c) \begin{pmatrix} {}^{n}Q_{1a} \\ {}^{n}Q_{2a} \\ {}^{n}Q_{3a} \\ {}^{n}Q_{4a} \end{pmatrix}, \tag{39}$$

where the matrix (c) is given by

$$(c) = (p)(a), \tag{40}$$

and

$$(p) = \begin{pmatrix} b_{11} & b_{12} & 0 & 0 \\ b_{21} & b_{22} & 0 & 0 \\ 0 & 0 & e^{-j\alpha} & 0 \\ 0 & 0 & 0 & e^{-j\mu} \end{pmatrix}. \tag{41}$$

Eq. (39) summarizes four simultaneous finite difference equations with constant coefficients. The solution is of the following form [19]:

$${}^{n+1}Q_{la} = e^{-\phi}\, {}^{n}Q_{la}, \tag{42}$$

where $l=1, 2, 3, 4$.

The parameter ϕ denotes the phase shift per unit cell in the presence of the electron beam, and is used in the calculation of gain. The values of ϕ are obtained by the solution of the following determinantal equation [38], which results when (39) and (42) are combined:

$$\left| (c) - e^{-\phi} \mathcal{I} \right| = 0, \tag{43}$$

where $\mathcal{I}$ is the unit or identity matrix [39] of order 4×4. Appendix II shows that (43) can be reduced to the following characteristic equation:

$$\begin{aligned} &\left\{ \left[a_{33} - e^{-(\phi - j\alpha)} \right] \left[a_{44} - e^{-(\phi - j\mu)} \right] + a_{34}{}^2 \right\} \\ &\quad \cdot \left\{ \cosh \phi - \cos \theta - j\Delta \hat{a}_{12} \sin \theta \right\} \\ &\quad = j \left\{ \hat{a}_{31}{}^2 \left[a_{44} - e^{-(\phi - j\mu)} \right] - \hat{a}_{41}{}^2 \left[a_{33} - e^{-(\phi - j\alpha)} \right] \right. \\ &\qquad \left. - 2a_{34}\hat{a}_{41}\hat{a}_{31} \right\} \sin \theta, \end{aligned} \tag{44}$$

where $\hat{a}_{13}$, $\hat{a}_{14}$, $\hat{a}_{31}$, $\hat{a}_{41}$ and $\Delta\hat{a}_{12}$ are listed in Appendix II, and are related to K_π, the midshunt characteristic impedance of the equivalent smooth transmission line shown in Fig. 3. Eq. (44) is the characteristic equation for the phase shift ϕ per period in the presence of the electron beam and is applicable to electrostatically and magnetically focused traveling-wave tubes.

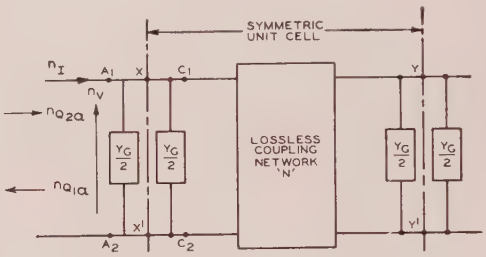

Fig. 3—Schematic diagram showing the reference terminals used to define the characteristic impedance of the equivalent smooth transmission line.

When (13), (14), (30), (32), (37), (38), (103)–(105), (110), (111), (137) and (139) are substituted in the characteristic equation (44), and higher-order terms in ω_a are neglected, the following equation results:

$$\begin{aligned} &\left[4\sinh^2\left(\frac{\phi - j\theta_e}{2} \right) + \frac{\omega_a{}^2 L}{u_a{}^2} (rg + d) \right] \\ &\quad \cdot \left[2 \sinh \left(\frac{\phi + j\theta}{2} \right) \sinh \left(\frac{\phi - j\theta}{2} \right) - j\Delta \hat{a}_{12} \sin \theta \right] \\ &\quad = -_2 |\hat{K}|^2 \left\{ \frac{\omega_a}{u_a} (rg + d) \right. \\ &\qquad \left. + j2\omega_a \tau_i \left[e^{(\phi - j\theta_e)} - 1 \right] \right\} \sin \theta. \end{aligned} \tag{45}$$

The preceding characteristic equation for ϕ applies under relatively small space-charge conditions, but is sufficiently accurate for analysis of typical traveling-wave tubes.

In (45), L, the period of the RF structure, is given by

$$L = d + g, \tag{46}$$

and θ_e, the dc electron transit angle per period, is given by

$$\theta_e = \frac{\omega}{u_a} \left[d + \frac{2g}{1 + \dfrac{u_b}{u_a}} \right]. \tag{47}$$

In this equation, u_b indicates the dc electron velocity corresponding to voltage V_b. In (45), $\hat{K}$ is given by

$$\hat{K} = \sqrt{\frac{K_\pi}{Z_0}}\, K_0. \tag{48}$$

Solution of the Characteristic Equation for Negligible Space-Charge Conditions

The following analysis concerns an RF circuit perturbed by a beam current so small that space-charge effects are negligible. If the dc electron transit angle per period θ_s is given by

$$\theta_e = \theta + 2m\pi, \tag{49}$$

where m denotes zero or a positive or negative integer, and if the phase shift per unit cell ϕ is represented by the expression

$$\phi = j\theta - \Delta\theta, \tag{50}$$

where $\Delta\theta$ is the perturbation caused by the beam current, the substitution of (49) and (50) in (45), with the assumption that $|\Delta\theta| \ll 1$, yields the equation

$$\Delta\theta^3 = -j2\,|\hat{K}|^2 \frac{\omega_a}{u_a}(rg + d). \tag{51}$$

The fourth-order equation (45) has been reduced to the third-order (51) by the preceding approximations. The same reduction has been performed by Pierce [40]. The three roots of (51) lead to a solution of the form

$${}^nQ_{\text{las}} = e^{-j n\theta_e} \sum_{k=1}^{3} {}^0Q_{\text{lak}} \cdot \exp\left(\delta_k \left\{\frac{2\,|\hat{K}|^2\omega_a(rg + d)}{u_a\theta_e^3}\right\}^{1/3} n\theta_e\right), \tag{52}$$

where δ_k denotes a cube root of $(-j)$.

According to Pierce's [41] notation, the parameter δ_1 gives rise to a component which increases exponentially, and which results in amplification. ${}^0Q_{\text{lak}}$ denotes the lth normalized RF variable associated with the kth cube root of $(-j)$, and ${}^nQ_{\text{las}}$ is the sum of the solutions resulting from the three roots of (51). A phase constant β_e is defined as follows:

$$\beta_e = \frac{\theta_e}{L}. \tag{53}$$

When the dc velocity of the electron beam is uniform $(u_a = u_b = u_0)$, the combination of (46), (47) and (53) gives the expression

$$\beta_e = \frac{\omega}{u_0}, \tag{54}$$

which agrees with Pierce's [42] definition of β_e. The combination of (52) and (53) gives the expression

$${}^nQ_{\text{ras}} = e^{-j\beta_e nL} \sum_{k=1}^{3} {}^0Q_{\text{lak}} \cdot \exp\left(\delta_k \left\{\frac{2\,|\hat{K}|^2\omega_a(rg + d)}{u_a\theta_e^3}\right\}^{1/3} \beta_e nL\right). \tag{55}$$

Eq. (55) is of the same form as Pierce's [14] (2.33), provided the length nL is interpreted as Pierce's coordinate z, and a generalized gain parameter $\tilde{C}$, which includes the periodic variations of dc electron velocity, is defined by the following equation:

$$\tilde{C} = \left\{\frac{2\,|\hat{K}|^2\omega_a(rg + d)}{u_a\theta_e^3}\right\}^{1/3}. \tag{56}$$

The quantity nL in (55) is not a continuous variable, and is significant only for integer values of n. It represents the length measured from the input of the zeroth interaction region to the input of the nth interaction region. When (48), (56), and (112) are combined, the following equation for the generalized gain parameter $\tilde{C}$ is obtained:

$$\tilde{C} = \left\{\frac{I_0 K_\pi u_a}{V_a(u_a + u_b)^2}\left[\frac{\sin\theta_0}{\theta_0} - \frac{1}{2}\left(1 - \frac{u_b}{u_a}\right)\frac{\sin^2(\theta_0/2)}{(\theta_0/2)^2}\right]^2 \frac{\omega(rg + d)}{\theta_e^3}\right\}^{1/3}, \tag{57}$$

where θ_0 is one half the gap transit angle. This element can be obtained through the combination of (46), (47), (101) and (107). Therefore,

$$\theta_0 = \frac{\theta_e g/L}{\left(1 + \frac{u_b}{u_a}\right) + \left(1 - \frac{u_b}{u_a}\right) g/L}. \tag{58}$$

It is now necessary to compare the value of $\tilde{C}$, given by (57), with that obtained from previous theories [13]–[15] for the special case of uniform beam velocity. When $u_b = u_a$, (46), (47), (49), (57), (58), and (106) can be combined to give the equation

$$C_u = \left(\frac{I_0 K}{4V_0}\right)^{1/3}, \tag{59}$$

where C_u denotes the value of $\tilde{C}$, and V_0 is the value of V_a when $u_a = u_b$. The parameter K is given by

$$K = \frac{K_\pi \sin^2\left[(\theta + 2m\pi)\frac{g}{2L}\right]}{(\theta + 2m\pi)^4\left(\frac{g}{2L}\right)^2}. \tag{60}$$

Using Pierce's notation [43], the impedance parameter K for the mth space harmonic, denoted by K_m, is given by

$$K_m = \frac{E_m^2}{2\beta_m^2 P}. \tag{61}$$

The combination of Pierce's [14] (2.28), (2.41), (2.42), (4.39), (4.61) and (4.67) with (59)–(61) yields the following expression:

$$C = \frac{H\left(\frac{b}{\rho}, \gamma_m\rho\right)}{I_0^{2/3}(\gamma_m\rho)} C_u, \tag{62}$$

This relationship is applicable to gridless structures if it is assumed that the correction factor $H(b/\rho, \gamma_m\rho)/I_0^{2/3}(\gamma_m\rho)$ is valid for both magnetically and electrostatically focused electron beams.

When (6), (46), (47), (49), (57), (59), (60), (62) and (63) are combined, and the previously mentioned conditions are imposed, the following expression for the normalized gain parameter C_N results:

$$C_N = \left\{ \frac{4\left[1 + (r-1)\frac{g}{L}\right]\left[\frac{\sin\theta_0}{\theta_0} - \frac{1}{2}\left(1 - \frac{u_b}{u_a}\right)\frac{\sin^2(\theta_0/2)}{(\theta_0/2)^2}\right]^2}{\left(1 + \frac{u_b}{u_a}\right)^2\left[1 + \left(\frac{1 - (u_b/u_a)}{1 + (u_b/u_a)}\right)\frac{g}{L}\right]^3 \frac{\sin^2(g\theta_e/2L)}{(g\theta_e/2L)^2}} \right\}^{1/3}. \tag{64}$$

where C denotes Pierce's gain parameter [44], $I_0(\gamma_m\rho)$ is the modified Bessel function of the first kind of order zero, $H(b/\rho, \gamma_m\rho)$ is Pierce's beam impedance reduction factor [45], and γ_m is given by Pierce's equation (4.63) [14].

The difference between C and C_u is caused by the one-dimensional model of the electron beam and RF circuit considered in this paper. Eq. (60) applies to a traveling-wave tube employing gridded gaps which produce a negligible variation of RF fields in the transverse direction. The gain parameter C, for a gridless structure, can be obtained by the multiplication of (59) by the correction factor $H(b/\rho, \gamma_m\rho)/I_0^{2/3}(\gamma_m\rho)$, as indicated by (62). For a typical traveling-wave tube ($b/\rho = 0.7$, $\gamma_m\rho = 1.4$) employing a round, solid electron beam, this correction factor is about 0.83.

The preceding discussion also indicates that the Llewellyn-Peterson equations predict gain when the dc transit angle per period θ_e differs from the cold phase shift per period θ by $2m\pi$, as shown by (49). The present theory predicts, therefore, the interaction between the electron beam and the space harmonics of the RF circuit [43]. For correct numerical values for the gain parameters associated with the space harmonics of a gridless structure, it is necessary to multiply the gain parameter obtained by use of Llewellyn-Peterson equations by the correction factor used in (62).

The gain parameter of the electrostatically focused tube will be compared with that of an equivalent magnetically focused tube, under the following conditions:

1) the characteristics of the RF circuits in the absence of the electron beam are identical;
2) the dc beam currents and dc current densities are equal;
3) the dc electron transit angles per period are equal.

If these conditions are satisfied, the performance of the electrostatically focused traveling-wave tube can be evaluated by means of a normalized gain parameter C_N, given by the following equation:

$$C_N = \frac{\tilde{C}}{C}. \tag{63}$$

For gridless structures, the dc velocity ratio u_b/u_a in (64) should be taken as the velocity ratio in the region of electron flow.

Eq. (64) gives the ratio between the gain parameter $\tilde{C}$ obtained with an electrostatically focused electron beam, and the gain parameter C obtained with a uniform dc velocity electron beam under conditions of negligible space charge, equal dc beam currents, dc current densities, and maximum gain.

Fig. 4 shows the normalized gain parameter C_N as a function of the dc velocity ratio u_b/u_a for three values of the gap-to-period ratio, and for a space-harmonic structure [9] having a dc transit angle per period of $3\pi/2$ radians. This figure shows that when the dc voltage of the drift tubes is lower than that of the focusing electrodes, the gain parameter realized by use of an electron beam having a periodically varying dc velocity is greater than that realized by use of a uniform dc velocity electron beam. The opposite is true for the reverse type of dc voltage distribution for practical gap-to-period and dc velocity ratios.

For a dc transit angle per period in the range from π to 2π radians, this difference between the gain parameters results mainly from the difference between the

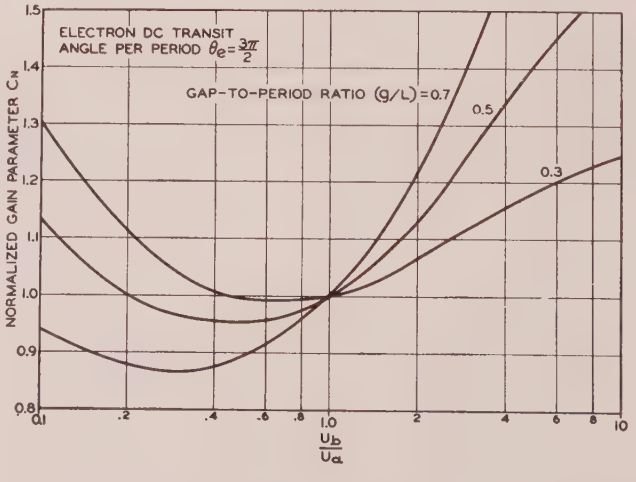

Fig. 4—Normalized gain parameter C_N as a function of dc-velocity ratio u_b/u_a, for several values of gap-to-period ratio and for a constant value of θ_e.

couplings of the electron beam to the RF circuit. For uniform dc velocity electron beams, this coupling is measured by means of the beam-coupling coefficient M, given by the following expression [46]:

$$M = \frac{\sin \theta_0'}{\theta_0'}, \tag{65}$$

where θ_0' is obtained from (58) by making u_b equal to u_a.

Eq. (64) shows that the beam-coupling coefficient $\tilde{M}$ for the electrostatically focused tube can be defined as follows:

$$\tilde{M} = \frac{\sin \theta_0}{\theta_0} - \frac{1}{2}\left(1 - \frac{u_b}{u_a}\right)\frac{\sin^2\left(\frac{\theta_0}{2}\right)}{\left(\frac{\theta_0}{2}\right)^2}, \tag{66}$$

where θ_0 is given by (58). When the dc transit angle per period θ_e is in the range from π to 2π radians, (65) and (66) yield the following inequalities:

$$|\tilde{M}| > |M| \quad \text{for} \quad u_b/u_a > 1, \tag{67}$$

$$|\tilde{M}| < |M| \quad \text{for} \quad u_b/u_a < 1. \tag{68}$$

This difference between the beam-coupling coefficients has a marked effect on the gain parameters of electrostatically focused traveling-wave tubes having practical gap-to-period and dc velocity ratios and dc transit angles per period in the range from π to 2π radians.

Fig. 5 shows the normalized gain parameter C_N as a function of the dc electron transit angle θ_e for two values of gap-to-period and dc velocity ratios for an electrostatically focused traveling-wave tube designed to operate in the range $\pi < \theta_e < 2\pi$. Because the dc transit angle θ_e is proportional to frequency, Fig. 5 also shows the variation of the normalized gain parameter with frequency at constant values of focusing voltages.

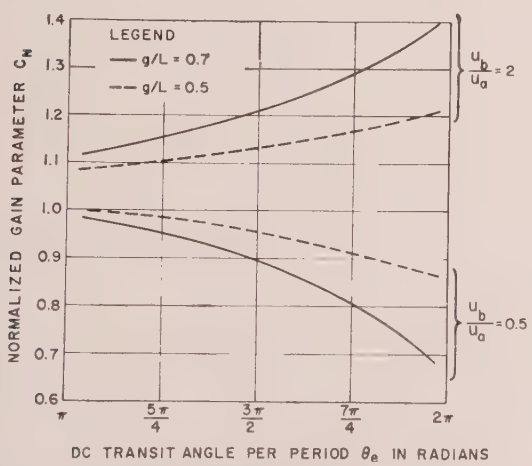

Fig. 5—Normalized gain parameter C_N as a function of dc transit angle per period θ_e for two values of gap-to-period and dc-velocity (u_b/u_a) ratios for a tube designed to operate in the range $\pi < \theta_e < 2\pi$.

Solution of the Characteristic Equation in Presence of Space-Charge

The following analysis concerns the interaction between the electron beam and the RF circuit under space-charge conditions. By analogy with Pierce's [47] analysis, parameters b and δ are introduced as shown in Table I. The interaction between the electron beam and

TABLE I
Pierce's and Analogous Parameters Used for Analysis

Pierce's Parameter		Analogous Parameter	
$\Gamma_1 = j\beta_e(1+Cb)$	(69)	$\theta = \theta_e(1+\tilde{C}b) - 2m\pi$	(70)
$\Gamma = j\beta_e - \beta_e C\delta$	(71)	$\phi = j\theta_e - \theta_e\tilde{C}\delta - j2m\pi$	(72)

the space harmonics of the RF circuit can be analyzed by substitution of integer values for m in (70) and (72). When (45), (56), (70) and (72) are combined with the conditions (73)–(76),

$$\theta_e\tilde{C}\,|\,\delta\,| \ll 1, \tag{73}$$

$$\theta_e\tilde{C}\,|\,\delta + jb\,| \ll 1, \tag{74}$$

$$\tilde{C}\,|\,b + j\delta\,| \ll 1, \tag{75}$$

$$\theta_e\tilde{C}\,|\,\delta + jb\,| \gg |\,\Delta\hat{a}_{12}\,|, \tag{76}$$

the parameter δ can be calculated from the following equation:

$$\delta^2 = \frac{1}{j\delta - b} - \frac{\omega_a{}^2 L(rg + d)}{\theta_e{}^2\tilde{C}^2 u_a{}^2}. \tag{77}$$

Eq. (57) shows that the gain parameter $\tilde{C}$ is proportional to $I_0{}^{1/3}$, and a detailed analysis [36] shows that $\Delta\hat{a}_{12}$ is approximately proportional to I_0. Condition (76) will hold, therefore, for small values of I_0.

If the space-charge parameter ($\tilde{Q}\tilde{C}$) of the electrostatically focused tube is defined by the following equation,

$$\tilde{Q}\tilde{C} = \frac{\omega_a{}^2 L(rg + d)}{4\theta_e{}^2\tilde{C}^2 u_a{}^2}, \tag{78}$$

then (77) and (78) can be combined to give

$$\delta^2 = \frac{1}{j\delta - b} - 4\tilde{Q}\tilde{C}, \tag{79}$$

which has the same form as Pierce's [48] equation for δ.

The generalized expression for the space-charge parameter ($\tilde{Q}\tilde{C}$) given by (78) includes the periodic variations of dc electron velocity. The space-charge parameter QC of a magnetically focused tube can be obtained from (78) if u_a is made equal to u_b. The parameter is then given by

$$QC = \frac{\omega_p{}^2}{4C^2\omega^2}, \tag{80}$$

where ω_p denotes the value of ω_a realized in the magnetically focused tube. Eq. (80) is the usual approximate expression for the space-charge parameter of a magnetically focused traveling-wave tube obtained by other methods of analysis [49]. If the transverse dimensions of the electron beam are finite, a plasma frequency reduction factor [23] must be used in conjunction with (80).

If the three conditions used for the definition of the normalized gain parameter C_N are satisfied, the performance of the electrostatically focused tube under space-charge conditions can be evaluated by means of a normalized space-charge parameter, $(QC)_N$, defined by the following equation:

$$(QC)_N = \frac{\tilde{Q}\tilde{C}}{QC} . \tag{81}$$

When (5), (47), (78), (80), and (81) are combined, and the conditions used for the definition of $(QC)_N$ are imposed, the normalized space-charge parameter $(QC)_N$ is given by

$$(QC)_N = \frac{1 + (r-1)\frac{g}{L}}{C_N^2 \left[1 + \left(\frac{1 - \frac{u_b}{u_a}}{1 + \frac{u_b}{u_a}} \right) \frac{g}{L} \right]^3} . \tag{82}$$

Eq. (82) gives the ratio between the space-charge parameter of an electrostatically focused tube and the space-charge parameter of an equivalent magnetically focused tube under conditions of equal dc beam currents, dc current densities, and, very nearly, maximum gain. Again, for gridless structures, the dc velocity ratio u_b/u_a in (82) should be taken as the velocity ratio in the region of electron flow.

Fig. 6 shows the normalized space-charge parameter as a function of the dc velocity ratio u_b/u_a for several values of the gap-to-period ratio, and for a constant value of θ_e. Fig. 7 shows the normalized space-charge parameter as a function of θ_e for two values of gap-to-period and dc velocity ratios for an electrostatically focused traveling-wave tube designed to operate in the range $\pi < \theta_e < 2\pi$. This figure also shows the variation of the normalized space-charge parameter with frequency at constant values of focusing voltages.

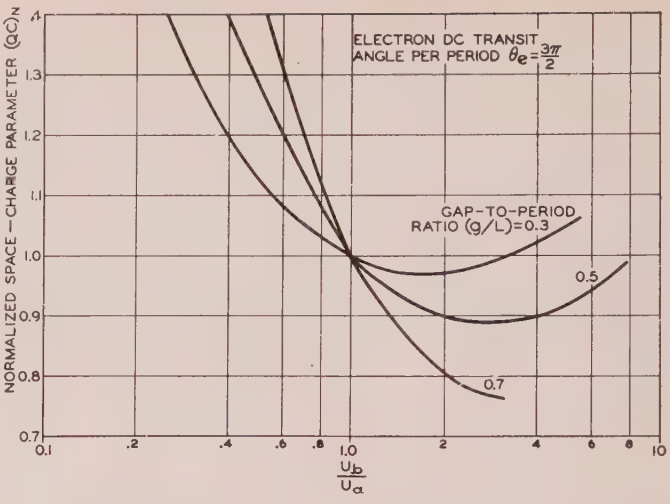

Fig. 6—Normalized space-charge parameter $(QC)_N$ as a function of dc-velocity ratio u_b/u_a, for several values of gap-to-period ratio and for a constant value of θ_e.

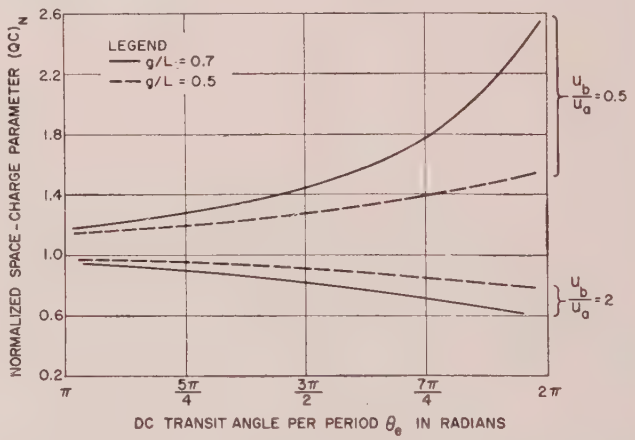

Fig. 7—Normalized space-charge parameter $(QC)_N$ as a function of dc transit angle per period θ_e for two values of gap-to-period and dc-velocity (u_b/u_a) ratios for a tube designed to operate in the range $\pi < \theta_e < 2\pi$.

RF Interaction Between a Sinusoidal Field and Electron Beam Having Uniform DC Velocity

To illustrate the generality of the present analysis, it will be shown briefly how the characteristic equation (45) can be applied to RF interaction between an electron beam moving at uniform dc velocity, and an RF electric field which, in the absence of electrons, varies as $e^{j\omega t - \Gamma_1 z}$, where Γ_1 is the propagation constant. Heffner's [15] treatment of this problem involves the subdivision of the tube into a large number of short cells, each cell consisting of an infinitely small interaction region ($g=0$), and a drift region sufficiently short to justify the following approximations:

$$e^{-\Gamma_1 d} \simeq 1 - \Gamma_1 d \tag{83}$$

$$e^{-\Gamma d} \simeq 1 - \Gamma d, \tag{84}$$

where Γ is the propagation constant in the presence of the electron beam. With the notation used in this paper,

$$\Gamma_1 d = j\theta \tag{85}$$

$$\Gamma d = \phi. \tag{86}$$

Substitution of (46)–(48), (53), (54), (83)–(86), (101), (107), and (112) in (45), with $u_a = u_b$ and $g = 0$, produces a characteristic equation for Γ identical with Heffner's [15] (58) which, in turn, is equivalent to Pierce's [14] (7.10).

Gain and Bandwidth of the Electrostatically Focused Traveling-Wave Tube

If the values of $(\tilde{Q}\tilde{C})$ and b are such that an exponentially increasing component exists, then combination of (42) and (72) shows that for large values of n, the RF quantities vary as follows:

$$\exp\left[-jn\theta_e(1 + j\tilde{C}\delta_1)\right] \tag{87}$$

where δ_1 denotes the root of the exponentially increasing component. Following Pierce's [50] notation, let

$$\delta_1 = x_1 + jy_1, \tag{88}$$

where x_1 and y_1 are real numbers. Let a wavelength λ_e be defined as follows:

$$\lambda_e = \frac{2\pi L}{\theta_e}. \tag{89}$$

Combination of (87)–(89) shows that for large values of n, the RF quantities vary as follows:

$$\exp(2\pi x_1\tilde{C}N)\exp\left[-j2\pi(1 - \tilde{C}y_1)N\right] \tag{90}$$

where

$$N = \frac{nL}{\lambda_e}. \tag{91}$$

Eq. (90) corresponds to Pierce's [14] (8.6).

The gain G of the traveling-wave tube is defined as follows:

$$G = 20\log_{10}\left|\frac{{}^nV}{{}^0V}\right|, \tag{92}$$

where ${}^nV$ denotes the RF gap voltage across the nth interaction region and ${}^0V$ is the input gap voltage.

The method of calculating the gain of the electrostatically focused traveling-wave tube is similar to that used by Pierce [51] and Heffner [52]. The detailed analysis [36] shows that the gain G of the electrostatically focused tube can be calculated from an expression having the same form as that derived for magnetically focused tubes [51], [53].

Therefore,

$$G = \tilde{A} + \tilde{B}\tilde{C}N, \tag{93}$$

where $\tilde{A}$ and $\tilde{B}$ denote, respectively, the initial loss and growing wave parameters for the electrostatically focused tube. The analysis presented in this paper gives the following expressions for these parameters:

$$\tilde{A} = 20\log_{10}\left|\frac{1 + \dfrac{4\tilde{Q}\tilde{C}}{\delta_1{}^2}}{1 + \left(\dfrac{\delta_2}{\delta_1}\right)^2\left(\dfrac{\delta_3 - \delta_1}{\delta_2 - \delta_3}\right) + \left(\dfrac{\delta_3}{\delta_1}\right)^2\left(\dfrac{\delta_1 - \delta_2}{\delta_2 - \delta_3}\right)}\right| \tag{94}$$

and

$$\tilde{B} = 40\pi(\log_{10} e)x_1. \tag{95}$$

For small values of $\tilde{C}$, (94) agrees with Birdsall's [54] (12) and (13). The deltas in (94) denote the three roots of (79).

When the values of $\tilde{C}$ and $\tilde{Q}\tilde{C}$, including the periodic variations of dc velocity are determined, numerical values for $\tilde{A}$ and $\tilde{B}$ can be obtained from graphical data available for traveling-wave tubes employing uniform dc velocity beams [14], [54]. The gain of the electrostatically focused traveling-wave tube can then be calculated from (93).

Application of Pierce's [55] results to Figs. 4–7 and (93), indicates that when the dc transit angle per period is in the range from π to 2π radians, and the dc voltage of the drift tubes is lower than that of the focusing electrodes, not only the gain, but also the bandwidth of an electrostatically focused space-harmonic structure are greater than the gain and bandwidth of the same structure employing a uniform dc velocity electron beam. The opposite is true for the reverse type of dc voltage distribution for practical gap-to-period and dc velocity ratios. The differences between the performances of the electrostatically focused tube have been verified experimentally.

Numerical Example

The gain of the electrostatically focused traveling-wave tube can be calculated most conveniently by use of the gain and space-charge parameters of an equivalent magnetically focused tube satisfying the criteria used for definition of the normalized gain parameter C_N, given by (64). The gain and space-charge parameters for the electrostatically focused tube can then be obtained from (64) and (82), respectively, or from Figs. 4–7. The gain of the electrostatically focused tube is then calculated in the same way as that used for magnetically focused tubes.

For this example, an equivalent magnetically focused traveling-wave tube having the following parameters is considered:

gain parameter C	$= 0.08$,
space-charge parameter QC	$= 0.25$,
gap-to-period ratio	$= 0.6$,
electron dc transit angle per period (θ_e) at maximum small-signal gain	$= 3\pi/2$,
number of cells	$= 20$.

From (89) and (91), $N=15$. From Pierce's [14] Figs. 8.10 and 9.4, $B=37$ and $A=-6.8$. The gain of the magnetically focused tube calculated from (93) is 37.6 db.

If the velocity ratio u_b/u_a in the region of the electron flow is $\sqrt{3}$, Fig. 4 and (63) yield a value for $\tilde{C}$ of 0.09. From Fig. 6 and (81), $(\tilde{Q}\tilde{C})=0.22$. From Pierce's [14] Figs. 8.10 and 9.4, $\tilde{B}=38$ and $\tilde{A}=-6.9$. From (93), the gain of the electrostatically focused traveling-wave tube is 44.5 db.

If the velocity ratio u_b/u_a in the region of electron flow is $1/\sqrt{3}$, $\tilde{C}=0.074$, and $(\tilde{Q}\tilde{C})=0.316$. From Pierce's [14] Figs. 8.10 and 9.4, $\tilde{B}=36$ and $\tilde{A}=-6.6$. From (93), the gain of the electrostatically focused traveling-wave tube is 33.4 db.

Experimental Results

The main objective of the initial RF tests was to evaluate the performance of the experimental electrostatically focused high-power traveling-wave tube, at small and large RF signal levels. In order to obtain a preliminary evaluation of the tube's performance, the initial RF tests were made prior to the development of an attenuator. In the absence of the attenuator, difficulties caused by oscillations were avoided by use of a short assembly consisting of only 11 cells. The tube employed gridless focusing electrodes and was designed for operation in the range $\pi<\theta_e<2\pi$. The solid curves in Fig. 8(a) show the measured maximum small-signal gain of this tube as a function of the cube root of the peak dc beam current for the two types of dc voltage distributions shown. The dashed curves were calculated from the expressions presented in this paper. For the calculation of the theoretical curves, the gain parameter C and the space-charge parameter QC realized at a given collector current in an equivalent magnetically focused tube were first calculated by the usual methods [13], [15]. The gain of the electrostatically focused tube was then calculated by use of the method described in the preceding section.

When the dc voltage of the drift tubes was lower than that of the focusing electrodes, the gain of the amplifier was greater than that measured for the reverse type of voltage distribution. The same difference in tube performance was measured at large RF signal levels. The theoretical and experimental results are, therefore, in agreement. Fig. 8(b) shows the voltages which were applied to the focusing electrodes for realization of maximum small-signal gain.

Conclusions

A theory of small-signal RF interaction between an electron beam having a periodically varying dc velocity and a traveling-wave circuit has been derived by the combination of the Llewellyn-Peterson equations with the space-charge wave theory and Pierce's theory for traveling-wave tubes. The limitations of the Llewellyn-

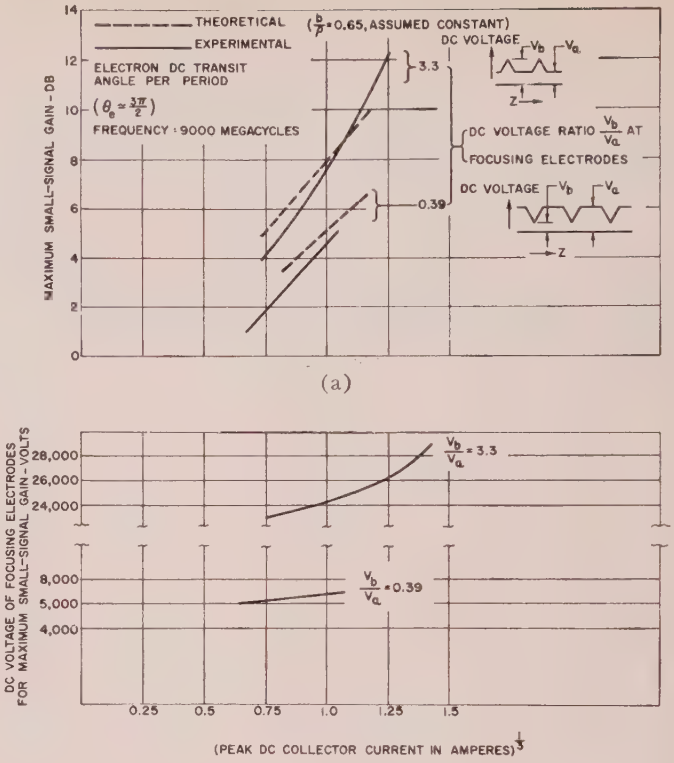

Fig. 8—(a) Theoretical and measured maximum small-signal gains of an experimental electrostatically focused high-power traveling-wave tube as a function of the cube root of peak dc collector current for two types of dc-voltage distributions. (b) DC voltage of focusing electrodes for maximum small-signal gain as a function of the cube root of peak dc collector current.

Peterson equations to one-dimensional problems has been overcome by the introduction of correction factors obtained from Pierce's theory of traveling-wave tubes.

The Llewellyn-Peterson equations were used to derive a characteristic equation for the phase shift per period of the traveling-wave circuit in the presence of an electron beam. This characteristic equation is applicable to magnetically focused traveling-wave tubes and a new type of electrostatically focused tube. Expressions for the gain and space-charge parameters of the electrostatically focused tube, which include the periodic variations of dc electron velocity, have been derived. When the dc electron velocity is uniform, as in a magnetically focused tube, these expressions reduce to the forms derived by Pierce. In addition to their suitability for analysis of electrostatically focused traveling-wave tubes, the Llewellyn-Peterson equations provide a method of analyzing magnetically focused traveling-wave tubes alternative to those used previously.

It was also shown that the gain of the electrostatically focused tube can be calculated by use of the tube's gain and space-charge parameters and of graphical data available for magnetically focused tubes.

Depending upon the voltage distribution in the region of electron flow, the gain and bandwidth of a traveling-

wave RF circuit employing a periodically varying dc velocity can be greater or smaller than the gain and bandwidth of the same RF circuit employing a uniform dc velocity electron beam.

Experimental results obtained on an electrostatically focused traveling-wave tube are in agreement with the theory.

The results presented in this paper show that the performance of traveling-wave tubes can be improved by use of a suitable periodic variation of dc electron velocity.

Appendix I

Derivation of Coefficients a_{31}–a_{44}

Consider region A_3A_4–B_1B_2 shown in Fig. 2(a). From the Llewellyn-Peterson equations [56]:

$$\begin{pmatrix} q_b \\ v_b \\ I \end{pmatrix} = (M_1) \begin{pmatrix} q_a \\ v_a \\ I \end{pmatrix}, \tag{96}$$

where

$$(M_1) = \begin{pmatrix} E_1^* & F_1^* & D_1^* \\ H_1^* & I_1^* & G_1^* \\ 0 & 0 & 1 \end{pmatrix}. \tag{97}$$

The coefficients E_1^*–G_1^* are identical with the E^*–G^* coefficients in the Llewellyn-Peterson [57] equations. The subscript 1 denotes reference to region A_3A_4–B_1B_2. A relationship similar to that given by (96) applies to region B_1B_2–C_3C_4. This relationship can be combined with (96) to give the following equation:

$$\begin{pmatrix} q_c \\ v_c \\ I \end{pmatrix} = (m) \begin{pmatrix} q_a \\ v_a \\ I \end{pmatrix}, \tag{98}$$

where

$$(m) = (M_2)(M_1) \tag{99}$$

The elements of (M_2) are obtained by interchanging the dc velocities u_a and u_b in the corresponding elements of (M_1).

When (5) is combined with Llewellyn-Peterson's [16] (2), and if it is assumed that the space-charge factor ζ is small, then

$$\omega_a \simeq \frac{\sqrt{2\zeta\left(1 + \dfrac{u_b}{u_a}\right)}}{T_0}, \tag{100}$$

where

$$T_0 = \frac{g}{u_a + u_b}. \tag{101}$$

If the current density I_D is uniform,

$$I_0 = AI_D, \tag{102}$$

where A is the cross-sectional area of the electron beam. When the total RFΔ current density I is zero, (3)–(8), (100)–(102), and appropriate elements of the (m) matrix can be combined to give (15) and (16). For small values of ζ, the coefficients a_{33}–a_{44} in (15) and (16) are given by

$$a_{33} = \left[1 - j\frac{\omega_a g}{2u_a}(1 + r) - \frac{\omega_a^2 g^2}{u_b(u_b + u_a)}\right] e^{-j2\theta_0} \tag{103}$$

$$a_{34} = -a_{33} = -j\frac{\omega_a g}{2u_a}(1 - r)e^{-j2\theta_0} \tag{104}$$

$$a_{44} = \left[1 + j\frac{\omega_a g}{2u_a}(1 + r) - \frac{\omega_a^2 g^2}{u_b(u_a + u_b)}\right] e^{-j2\theta_0}, \tag{105}$$

where r is the dc velocity variation coefficient given by

$$r = \frac{2}{\dfrac{u_b}{u_a}\left(1 + \dfrac{u_b}{u_a}\right)}, \tag{106}$$

and θ_0 is one half the gap transit angle given by the following expression:

$$\theta_0 = \omega T_0 \tag{107}$$

If the initial beam modulation q_a and v_a is zero, (98) gives the expressions

$$q_c = m_{13}I \tag{108}$$

$$v_c = m_{23}I. \tag{109}$$

Eqs. (20) and (21) and the coefficients a_{31}, a_{32}, a_{41}, a_{42} can be obtained by the combination of (1)–(8), (19), (102), (108), (109) and appropriate elements of the (m) matrix. For small values of ζ, the following relationships are obtained:

$$a_{31} = a_{32} = -K_0\left(1 - \omega_a\tau_i - j\frac{\omega_a g r}{2u_a}\right) \tag{110}$$

$$a_{41} = a_{42} = K_0\left(1 + \omega_a\tau_i + j\frac{\omega_a g r}{2u_a}\right), \tag{111}$$

where

$$K_0 = \sqrt{\frac{Z_0 I_0 \omega}{2V_a\omega_a}} \cdot \frac{u_a}{(u_a + u_b)}\left[\frac{\sin\theta_0}{\theta_0} - 2\left(1 - \frac{u_b}{u_a}\right)\frac{\sin^2(\theta_0/2)}{\theta_0^2}\right] e^{-j\theta_0}. \tag{112}$$

Appendix II

Derivation of Characteristic Equation for the Phase Shift Per Unit Cell in the Presence of an Electron Beam

The combination of (25), (26), (33), (40), and (41) produces the following series of equations:

$$c_{11} = b_{11} + a_{12}(b_{11} - b_{12}) \tag{113}$$

$$c_{12} = b_{12} + a_{12}(b_{11} - b_{12}) \tag{114}$$

$$c_{21} = b_{21} + a_{12}(b_{21} - b_{22}) \tag{115}$$

$$c_{22} = b_{22} + a_{12}(b_{21} - b_{22}). \tag{116}$$

From (113) and (114),

$$b_{11} - b_{12} = c_{11} - c_{12}. \tag{117}$$

From (115) and (116),

$$b_{22} - b_{21} = c_{22} - c_{21}. \tag{118}$$

It has been shown by Pierce [37] that for a lossless and reciprocal transducer,

$$|b| = 1. \tag{119}$$

The combination of (113)–(116) and (119) gives the expression

$$c_{11}c_{22} - c_{12}c_{21} = 1. \tag{120}$$

When (43) is solved by the usual method [38], and when (27), (28), (30), (32), (40), (41), (104), (117), (118), and (120) are combined, the characteristic equation for ϕ is as follows:

$$\begin{aligned}&\{[a_{33} - e^{-(\phi - j\alpha)}][a_{44} - e^{-(\phi - j\mu)}] + a_{34}^2\} \\ &\quad \cdot \{2\cosh\phi - (c_{11} + c_{22})\} \\ &= \{a_{31}^2[a_{44} - e^{-(\phi - j\mu)}] - a_{41}^2[a_{33} - e^{-(\phi - j\alpha)}] \\ &\qquad - 2a_{34}a_{41}a_{31}\}(c_{11} + c_{21} - c_{22} - c_{12}).\end{aligned} \tag{121}$$

Consider, initially, the periodically loaded transmission line in the absence of the electron beam. Let matrices (a') and (c'), associated with the RF circuit in the absence of the electron beam, be defined as follows:

$$\begin{pmatrix} {}^nQ_{1c} \\ {}^nQ_{2c} \end{pmatrix} = (a') \begin{pmatrix} {}^nQ_{1a} \\ {}^nQ_{2a} \end{pmatrix} \tag{122}$$

$$\begin{pmatrix} {}^{n+1}Q_{1a} \\ {}^{n+1}Q_{2a} \end{pmatrix} = (c') \begin{pmatrix} {}^nQ_{1a} \\ {}^nQ_{2a} \end{pmatrix}, \tag{123}$$

where

$$(c') = (b)(a'). \tag{124}$$

If the RF losses are zero [37],

$$|c'| = 1. \tag{125}$$

The phase shift per period [58] θ' of the equivalent smooth transmission line is given by

$$\frac{{}^{n+1}V}{{}^nV} = \frac{{}^{n+1}Q_{1a}}{{}^nQ_{1a}} = \frac{{}^{n+1}Q_{2a}}{{}^nQ_{2a}} = e^{-j\theta'}, \tag{126}$$

where ${}^nV$ and ${}^{n+1}V$ denote the RF voltages across the nth and $(n+1)$th gaps, respectively. The combination of (123), (125) and (126) yields the expression

$$\cos\theta' = \tfrac{1}{2}(c_{11}' + c_{22}'). \tag{127}$$

If there is no attenuation, (127) determines two real values of θ' of equal magnitudes but opposite signs [59]. From (1) and (2), and from Fig. 3, the peak RF current ${}^nI$, entering the input of the nth interaction region, is given by

$${}^nI = \sqrt{\frac{2}{Z_0}}\,({}^nQ_{2a} - {}^nQ_{1a}), \tag{128}$$

if

$${}^nV = {}^nV_{1a} + {}^nV_{2a} \tag{129}$$

and

$${}^{n+1}V = {}^{n+1}V_{1a} + {}^{n+1}V_{2a}. \tag{130}$$

The combination of (1), (2), (123), and (128)–(130) produces the following expression for ${}^{n+1}V$:

$$\begin{aligned}{}^{n+1}V &= \tfrac{1}{2}(c_{11}' + c_{21}' + c_{22}' + c_{12}'){}^nV \\ &\quad + \frac{Z_0}{2}(c_{22}' + c_{12}' - c_{11}' - c_{21}'){}^nI.\end{aligned} \tag{131}$$

The characteristic impedance $\hat{Z}$ of the equivalent smooth transmission line is given by

$$\frac{{}^nV}{{}^nI} = \hat{Z}. \tag{132}$$

Eqs. (1), (2), (113)–(116), (124), (126), (131) and (132) combine to give the expression

$$c_{11} + c_{21} - c_{22} - c_{12} = \frac{\hat{Z}_{1,2}}{Z_0}(c_{21}' + c_{12}' \pm j2\sin\theta), \tag{133}$$

where θ denotes the positive value of θ', and where $\hat{Z}_1$ and $\hat{Z}_2$ are the characteristic impedances of the equivalent smooth transmission line associated with the positive and negative values of θ', respectively [60]. From (133),

$$c_{11} + c_{21} - c_{22} - c_{12} = j\frac{2\hat{Z}_c}{Z_0}\sin\theta, \tag{134}$$

where

$$\hat{Z}_c = \frac{2\hat{Z}_1\hat{Z}_2}{\hat{Z}_2 - \hat{Z}_1}; \tag{135}$$

$\hat{Z}_c$ is equal to twice the impedance produced by the connection of $\hat{Z}_1$ and $\hat{Z}_2$ in parallel. Impedances $\hat{Z}_1$ and $\hat{Z}_2$ are referred to terminals A_1A_2 shown in Fig. 3. If the unit cell XY, shown in Fig. 3, is symmetric, then, from elementary circuit theory [59]

$$\hat{Z}_c = K_\pi, \tag{136}$$

where K_π is referred to terminals XX' shown in Fig. 3, and is the midshunt characteristic impedance [61] of the equivalent smooth transmission line. The coefficients associated with such a line are defined as follows:

$$\hat{a}_{31} = \sqrt{\frac{K_\pi}{Z_0}}\, a_{31} \tag{137}$$

$$\hat{a}_{13} = \sqrt{\frac{K_\pi}{Z_0}}\, a_{13} \tag{138}$$

$$\hat{a}_{41} = \sqrt{\frac{K_\pi}{Z_0}}\, a_{41} \tag{139}$$

$$\hat{a}_{14} = \sqrt{\frac{K_\pi}{Z_0}}\, a_{14}. \tag{140}$$

From (34), (113)–(118), (124), (127) and (134):

$$c_{11} + c_{22} = 2\cos\theta + j2\Delta\hat{a}_{12}\sin\theta, \tag{141}$$

where

$$\Delta\hat{a}_{12} = \frac{Y_B K_\pi}{2}. \tag{142}$$

Eq. (44) is obtained by combining equations (121), (134), and (137)–(141).

Acknowledgment

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Correspondence

WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table below.

The characteristics of the USFS, and its relation to time scales such as ET and UT2, have been described in a previous issue,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960. The last time adjustment was a retardation adjustment of 0.02 s on December 16, 1959.

WWV Frequency
With Respect to U. S. Frequency Standard

1960 August 1600 UT	Parts in 10^{10}†
1	−147
2	−146
3	−146
4	−146
5	−146
6	−146
7	−146
8	−145
9	−145
10	−145
11	−145
12	−145
13	−145
14	−145
15	−145
16	−145
17	−145
18	−145
19	−145
20	−145
21	−145
22	−146
23	−146
24	−146
25	−146
26	−147
27	−147
28	−147
29	−147
30	−148
31	−148

† A minus sign indicates that the broadcast frequency was low.

* Received by the IRE, September 26, 1960.

[1] "United States National Standards of Time and Frequency," Proc. IRE, vol. 48, pp. 105–106; January, 1960.

National Bureau of Standards
Boulder, Colo.

Esaki Diodes as Super-Regenerative Detectors*

A super-regenerative amplifier is an oscillator whose oscillations are allowed to build up and decay periodically. The periodic voltage or current which controls the initiation and interruption of the oscillations is obtained from a separate source, the quench oscillator, although the super-regenerative amplifier may also be made self-quenching. When oscillations are allowed to start, they build up from the signal whenever there is one present if it is at a higher level than the noise; otherwise, they build up from the noise. In the linear mode, the oscillations are quenched before they reach their equilibrium amplitude and their peak value is approximately proportional to the amplitude of the signal or noise wherefrom they started.

With the advent of tunnel diodes and the simplicity involved in their circuitry, super-regeneration may become again attractive for amplification and detection. Fig. 1 shows the schematic diagram of a simple super-regenerative detector using an Esaki diode. The tunnel diode is biased in the negative resistance region and the amplitude of the quench voltage is selected so as to move the operating point beyond the valley where a positive resistance exists and thus causes the oscillations to decay.

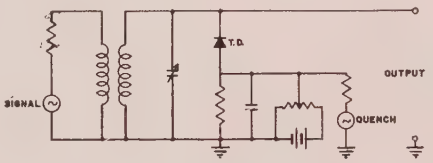

Fig. 1—Schematic diagram of super-regenerative detector circuit.

Fig. 2 shows oscillograms of the oscillations for different signal levels using a sinusoidal quench which is also shown in the photograph. The oscillator frequency is 6.25 Mc and the quench frequency is 100 kc.

Fig. 3 shows the oscillations obtained using a modulated signal (50 per cent modulation, 400-c modulation frequency) for different carrier frequencies in the vicinity of the oscillator frequency (6.25 Mc). Notice the time calibration in (a) is not the same as that in the rest of the oscillograms.

The units used in the experiments above were tunnel diodes with peak current = 0.6 ma, valley current = 0.06 ma, and a capacitance measured at the valley point ≃100 $\mu\mu$f.

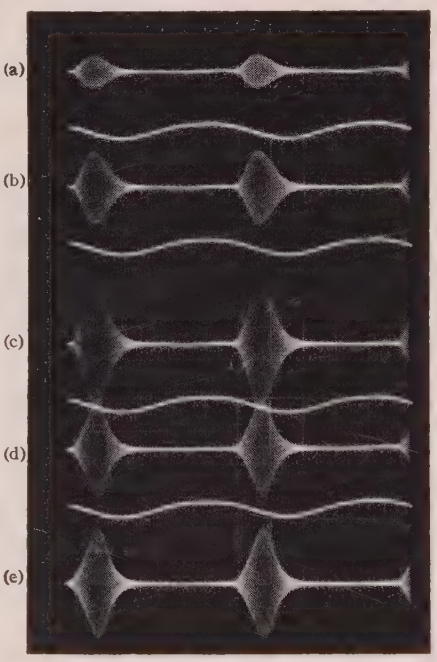

Fig. 2—Oscillograms of the oscillations for different signal levels using a sinusoidal quench voltage of 100 kc. (a) 22 mv signal; (b) 40 mv; (c) 60 mv; (d) 80 mv; (e) 100 mv. Horizontal calibration, 2 μsec/cm. Vertical calibration, (a)–(c), 0.05 volt/cm; (d)–(e), 0.1 volt/cm. Vertical calibration for quench voltage, 0.5 volt/cm.

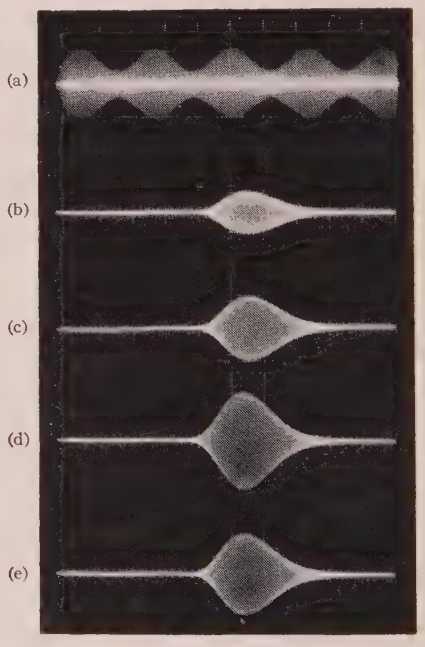

Fig. 3—(a) AM signal (carrier frequency, 6.25 Mc; modulation frequency, 400 c; 50 per cent modulation). (b) Oscillations with AM signal of 6 Mc carrier frequency. (c) Oscillation with AM signal of 6.1 Mc carrier frequency. (d) 6.25 carrier frequency. (e) 6.45 carrier frequency. Vertical calibration: (a)–(e), 0.05 volt/cm. Horizontal calibration: (a), 1 msec/cm; (b)–(e), 1 sec/cm.

A paper with a detailed analysis of Esaki diode super-regenerative detectors and experimental results is forthcoming.

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Theoretical Justification for Shot-Noise Smoothing in the Esaki Diode*

The Esaki diode is generally believed to make full shot noise in the negative conductance region[1-3] according to

$$\overline{i^2} = 2e\Delta f(I_{c\to v} + I_{v\to c}), \quad (1)$$

where e is the electronic charge, Δf is the width of a rectangular shaped incremental pass band, $I_{c\to v}$ is the long time average current caused by n-region conduction electrons tunneling into p-region valence states, and $I_{v\to c}$ is the long time average current caused by p-region valence electrons tunneling into n-region conduction states. The net direct current is

$$I = I_{c\to v} - I_{v\to c}. \quad (2)$$

The purpose of this note is to advance a crude theory of shot-noise smoothing in the negative conductance region to explain certain experimental observations. Stated briefly, the theory to be described predicts that most of the eligible electrons do not have infinitesimal probability of tunneling and, therefore, the diode need not be as noisy as was first assumed.

The current $I_{v\to c}$ is not smoothed because it comes from a small probability mechanism. Its shot noise should agree with the second component of (1).

The first component of (1) is reduced by smoothing, and the total mean-square noise current should be

$$\overline{i^2} = 2e\Delta f(\gamma^2 I_{c\to v} + I_{v\to c}), \quad (3)$$

where γ^2 is a smoothing factor roughly equal to that assumed by Hines.[4] The following discussion attempts to find a rough value for γ^2.

Using Esaki's notation,[5]

$$I_{c\to v} = A\int_{E_c}^{E_v} \rho_c(E)f_c(E)Z_{c\to v}f_v'(E)\rho_v(E)dE, \quad (4)$$

where

$\rho_c(E)$ = n-conduction band energy level density,
$\rho_v(E)$ = p-valence band energy level density,
$f_c(E)$ = probability of n-conduction state occupancy,
$f_v'(E) = 1 - f_v(E)$, where
$f_v(E)$ = probability of p-valence state occupancy,
$Z_{c\to v}$ = probability of penetrating the gap,
E = energy level,
E_c = lowest energy level in n-conduction band,
E_v = highest energy level in p-valence band, and
A = a constant multiplier.

In a differential energy range dE, the differential contribution to $I_{c\to v}$ is

$$dI_{c\to v} = A\rho_c(E)\rho_v(E)Z_{c\to v}f_c(E)f_v'(E)dE. \quad (5)$$

The last three factors of (5), namely

$$Z_{c\to v}f_c(E)f_v'(E), \quad (6)$$

form the joint probability of $c\to v$ tunneling in the energy range dE. This probability is important because the differential contribution to mean-square noise current is

$$2e\Delta f[1 - Z_{c\to v}f_c(E)f_v'(E)]dI_{c\to v}. \quad (7)$$

This formula for shot noise is easily derived, if we assume a binomial distribution for the total number of $c\to v$ electrons during any one sampling interval Δt, where

$$\Delta t = \frac{1}{2\Delta f}. \quad (8)$$

See Van der Ziel[6] for the mean-square deviation of a binomial distribution.

As forward bias is applied to the diode, the interval between E_c and E_v decreases. Fig. 2d of Lesk and Suran[1] shows that $f_c(E)$ and $f_v'(E)$ are almost unity in this energy range; making the joint probability (6) almost equal to $Z_{c\to v}$.

The probability $Z_{c\to v}$ might have a value of about 0.5 from the following hypothesis. Since both sides of the junction must be degenerate, the conduction electrons on the n side are free of the donor atoms, and move about like gas particles, subject only to scattering by impurity centers.[7] The lowest-energy electrons are responsible for the tunneling. Since tunneling requires no change of energy, these electrons are unaware of the potential hill created by the diffusion of the high-energy electrons and holes. This means that the electrons which tunnel are equally likely to move in any direction.

Those electrons on the edge of the junction have a 50 per cent probability of heading across a thin junction, and an equal probability of heading back deeper into the n material. This suggests a value of 0.5 for Esaki's $Z_{c\to v}$. This hypothesis is admittedly crude, and it is unable to take care of refinements such as the contribution from electrons deeper in the n material. It is believed, however, that because of the heavy doping, the electrons adjacent to the junction shield the others so that the value of 0.5 might be fairly universal for any electron which can tunnel.

If all the factors in (6) are approximately constant in the range of E from E_c to E_v, the integration of (7) is simplified because

$$1 - Z_{c\to v}f_cf_v' \quad (9)$$

can be taken outside the integral where it simply becomes γ^2, leading to the first term in (3). Since the first term accounts for most of the current, (2), and most of the noise, (3), the total shot noise will be reduced by approximately γ^2, which might be as low as 0.5 from the foregoing discussion.

In conclusion, it should be noted that not all tunnel diodes should show this noise reduction. It might be significant that the experiments of Chang, *et al.*,[8] use diodes having fairly large peak currents, indicating a large ratio of junction area to periphery. This could reduce the surface complications and improve the smoothing.

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* Received by the IRE, July 29, 1960.

[1] I. A. Lesk and J. J. Suran, "Tunnel diode operation and application," *Elec. Engrg.*, vol. 79, pp. 270–277; April, 1960.

[2] R. N. Hall, "Tunnel diodes," IRE Trans. on Electron Devices, vol. ED-7, pp. 1–9; January, 1960.

[3] D. I. Breitzer, "Noise figure of a tunnel diode mixer," Proc. IRE, vol. 48, pp. 930–936; May, 1960.

[4] M. E. Hines, "High frequency negative-resistance circuit principles for Esaki diode applications," *Bell Sys. Tech. J.*, vol. 39, pp. 477–513; May, 1960.

[5] L. Esaki, "New phenomenon in narrow germanium junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January 15, 1958.

[6] A. van der Ziel, "Fluctuation Phenomena in Semi-Conductors," Academic Press, Inc., New York, N. Y., pp. 13-14; 1959.

[7] William Shockley, "Electrons and Holes in Semiconductors," D. Van Nostrand Co., Inc., New York, N. Y., p. 283; 1950.

[8] K. K. N. Chang, G. H. Heilmeier and H. J. Prager, "Low noise tunnel diode down converter having conversion gain," Proc. IRE, vol. 48, pp. 854–858; May, 1960.

Noise Performance of Tunnel Diodes*

It is well known that the negative resistance element of a tunnel diode exhibits shot noise.[1] A tunnel diode also has a parasitic series resistance. If it is assumed that this exhibits thermal noise, some significant conclusions can be drawn. The relative contribution of these two noise sources to the diode's noise performance is frequency dependent. At frequencies much lower than the cutoff frequency (the frequency above which the diode's terminal impedance no longer exhibits a negative real part) the shot noise is normally the main contributor. At frequencies approaching the cutoff frequency the thermal noise becomes relatively more important and the tunnel diode's noise performance deteriorates.

This performance is determined from the noise equivalent circuit, Fig. 1. Here R_d is the diode negative resistance, R_s is the parasitic series resistance, and i_d and e_s are, respectively, the shot and thermal noise generators where

$$\overline{i_d^2} = 2eI_0\Delta f$$

$$\overline{e_s^2} = 4kTR_s\Delta f,$$

and I_0 is the dc current.

For mathematical convenience the shot noise is represented by the equivalent noise conductance G_e where

$$G_e = \frac{eI_0}{2kT}$$

$$\overline{i_d^2} = 4kTG_d\left(\frac{G_e}{G_d}\right)\Delta f$$

$$G_d = \frac{1}{R_d}.$$

For calculations of spot noise figure the circuit is put into the form of a pure series or shunt equivalent, Figs. 2 and 3, respectively. In Fig. 2, the negative real part of the im-

* Received by the IRE, May 31, 1960.

[1] J. J. Tiemann, "Tunnel Diodes and Their Use as Multifunctional Circuit Elements," presented at 1960 Internatl. Solid-State Circuits Conf., Philadelphia, Pa.; February 10, 1960.

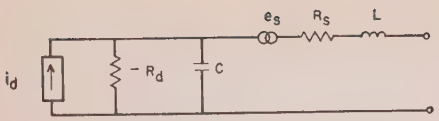

Fig. 1—Noise equivalent circuit.

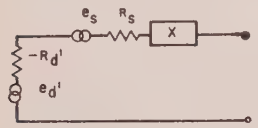

Fig. 2—Single frequency series noise equivalent.

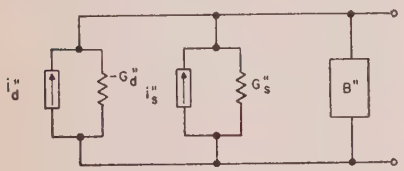

Fig. 3—Single frequency shunt noise equivalent circuit

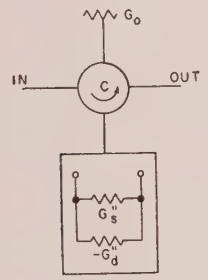

Fig. 4—Tunnel diode amplifier employing circulator.

pedance is reduced by the shunt capacity to the value

$$R_{d'} = \frac{R_d}{1 + \omega^2 R_d^2 C^2}. \tag{2}$$

The reactance is given by

$$jX = j\omega\left[L - \frac{CR_d^2}{1 + \omega^2 R_d^2 C^2}\right].$$

The noise in $R_{d'}$ is represented by $e_{d'}$ where

$$\overline{e_{d'}^2} = 4kTR_{d'}\left(\frac{G_e}{G_d}\right)\Delta f.$$

The noise in R_s is of course the same as given previously.

Transforming Fig. 2 into a parallel equivalent gives Fig. 3 where $R_{d'}$ and $e_{d'}$ transform into $G_{d''}$ and $i_{d''}$; R_s and e_s, into $G_{s''}$ and $i_{s''}$; and X into B''. The transforming equations are

$$G_{d''} = \frac{R_{d'}}{(R_s - R_{d'})^2 + X^2} \tag{3}$$

$$G_{s''} = \frac{R_s}{(R_s - R_{d'})^2 + X^2} \tag{4}$$

$$B'' = -\frac{X}{(R_s - R_{d'})^2 + X^2}$$

$$\overline{i_{d''}^2} = 4kT\left(\frac{G_e}{G_d}\right)G_{d''}\Delta f$$

$$\overline{i_{s''}^2} = 4kTG_{s''}\Delta f. \tag{5}$$

In each case, the noise in the negative real part of the immittance may be considered mathematically equivalent to thermal noise multiplied by the G_e/G_d ratio.

Comparing Fig. 3 with the models used by Chang,[2] and Hines and Anderson[3] the following equivalences are obtained:

$$G_{d''} = G$$
$$G_{s''} = G_1$$
$$\frac{G_e}{G_d} = \frac{eI_0}{2kTG}.$$

In Chang's paper G_1 apparently represented a parasitic circuit loss; here it represents the parasitic diode loss. In the previous papers, the negative conductance element was assumed to have a shot noise determined by the dc current. However, this must be modified to include the effects of the other circuit elements. Using (1) and (5), the following result is obtained:

$$\overline{i_{d''}^2} = 2eI_0\frac{G_{d''}}{G_d}\Delta f.$$

A negative resistance amplifier can be made unilateral by means of a circulator, Fig. 4. This arrangement also achieves the lowest possible noise measure.[4] The gain becomes high when the negative admittance approaches the characteristic admittance of the circulator. Assuming the signal source admittance G_g and circulator characteristic admittance G_0 are real and equal, and the tunnel diode reactance B'' is tuned out, this gives

$$G_g = G_0 = G_{d''} - G_{s''} + \epsilon$$

where ϵ is an arbitrarily small positive number. In the imit, as ϵ approaches zero, the gain approaches infinity and the noise figure becomes[5]

$$F = 1 + \frac{T\left(\frac{G_e}{G_d} + \frac{G_{s''}}{G_{d''}}\right)}{T_0\left(1 - \frac{G_{s''}}{G_{d''}}\right)} \tag{6}$$

where T_0 is a reference temperature. Eq. (6) is also significant to two-port tunnel diode preamplifiers. It can be shown that (6) is the minimum noise figure obtainable with such a preamplifier when it is followed by additional relatively noisy circuits. If it is assumed that $T = T_0$, (6) becomes a function only of the tunnel diode. It then is a very useful index of the diode's noise performance.

Assuming $T = T_0$ and making use of (2)–(5),

$$F = \frac{\left(1 + \frac{G_e}{G_d}\right)}{\left(1 - \frac{R_s}{R_d}\right)\left[1 - \left(\frac{f}{f_c}\right)^2\right]},$$

where f_c is the cutoff frequency given by

$$f_c = \frac{1}{2\pi CR_d}\sqrt{\frac{R_d}{R_s} - 1}.$$

The noise figure may be considered a function of three ratios, G_e/G_d, R_s/R_d, and f/f_c. The first assesses the shot noise relative to the negative conductance; the second, the parasitic series resistance relative to the negative resistance; and the third, the operating frequency relative to the diode's cutoff frequency. With germanium tunnel diodes $R_s/R_d \approx 0.02$ and $G_e/G_d \approx 1$. Thus at frequencies much smaller than f_c, a noise figure of 3.1 db is obtained; and this is almost entirely due to the diode shot noise. As the operating frequency approaches f_c the noise figure deteriorates rapidly. For example, the increase is 3 db for $f = 0.707\, f_c$, and 7.2 db for $f = 0.9\, f_c$. The frequency function is very important in microwave amplifiers.

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[2] K. K. N. Chang, "Low-noise tunnel-diode amplifier," Proc. IRE, vol. 47, pp. 1268–1269; July, 1959.
[3] M. E. Hines and W. W. Anderson, "Noise performance theory of Esaki (tunnel) diode amplifiers," Proc. IRE, vol. 48, p. 789; April, 1960.
[4] H. A. Haus and R. B. Adler, "Circuit Theory of Noisy Networks," John Wiley and Sons, Inc., New York, N. Y.; pp. 68–71; 1959.
[5] Hines and Anderson, *op. cit.*, see (5).

Resonant Modes in an Optical Maser*

Dual mirror interferometers such as Fabry-Pérot interferometers have been proposed to serve as resonators in optical masers.[1] It is common practice to consider that a Fabry-Pérot interferometer is simultaneously resonant for uniform plane waves traveling parallel to the axis and at certain discrete angles from the axis. It is clear, however, that diffraction will cause loss of power from the edges of the wavefront, and that this upon repeated reflection might cause a marked departure from uniform amplitude distribution across the mirror. When the interferometer is continuously supplied with uniform plane wave power from an external source, the internal distribution may depart from uniformity by a trivial amount. However, if the interferometer were filled with an active medium as required for an oscillating maser, a wave, once initiated, would be perpetuated by internal amplification and might be expected to exhibit the redistribution suggested.

In order to investigate this possibility, an IBM 704 computer was programmed to perform the experiment of bouncing a wave back and forth repeatedly between plane parallel mirrors of zero reflection loss. A uniform plane wave at one mirror was broken up into 200 Huygens' sources across the diameter, and their contributions to the field at the opposite mirror were summed at 200 points across the latter mirror. Simplification of the problem was achieved by assuming that the separation between mirrors

* Received by the IRE, July 11, 1960.
[1] A. L. Schawlow and C. H. Townes, "Infrared and optical masers," *Phys. Rev.*, vol. 112, pp. 1940–1949; December, 1958.

was large relative to their diameter, and that the diameter was large relative to a wavelength. Further, it was assumed that E and H are everywhere almost parallel to the mirrors. The resulting field distribution across the second mirror was used as a new initial wave and the resulting field was again computed. By repeating this computation hundreds of times it was observed that the diffraction loss not only caused the wave to grow weaker with successive transits, but did in fact cause the amplitude at the edges to drop to a small fraction of the center value. Moreover, after a sufficient number of transits the distribution converged to a steady state, an example of which is shown in Fig. 1, for the case of circular mirrors with

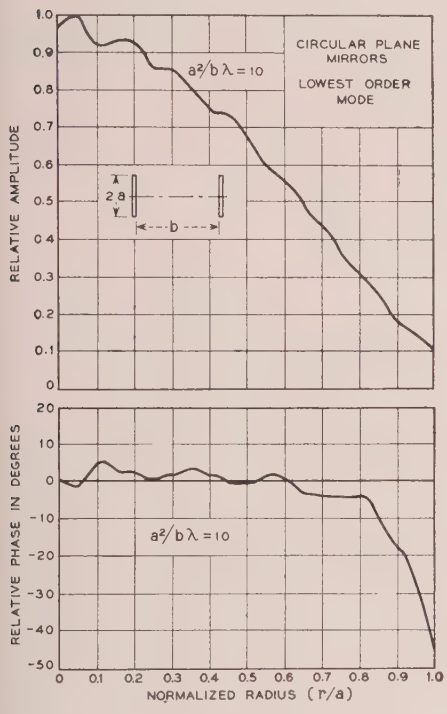

Fig. 1—Amplitude and phase distribution of electric field intensity of the lowest order mode for circular plane mirrors ($a = 0.316$ cm, $b = 100$ cm, $\lambda = 10^{-4}$ cm).

the dimensions indicated. These curves are plotted for 100 computed points along the radius. The wave represented by this distribution must be considered a normal mode of the interferometer. It has been shown that there are many normal modes representing higher orders of variation of field across the plates. Each of these has a characteristic loss and phase shift per bounce, analogous to the attenuation and phase constants for modes of propagation of a waveguide. The mode shown in Fig. 1 is the lowest order mode and has uniform polarization over the mirror.

In the light of these findings, uniform plane waves can no longer be regarded as stable modes for a Fabry-Pérot interferometer. A uniform plane wave must be made up of a large number of normal modes, each of which resonates for a slightly different plate spacing. The fact that such a decomposition is not normally observed in an optical interferometer can be predicted, knowing the mode separations and the resolving power of such instruments. Thus optical reflectivities are so poor that the line width of the best interferometer would contain hundreds of normal mode resonances. In the microwave range it might be expected that some of these resonances should be resolved.

In the case of an optical maser, which starts to build up oscillation from thermal noise excitation, the kind of wave which will enjoy the most rapid growth can be predicted using the normal mode picture. With the proper mirror spacing, the mode which should win the race to saturation would be the one with the lowest losses, namely, the lowest order symmetric mode shown in Fig. 1. The onset of saturation would naturally tend to level off the amplitude distribution shown.

The diffraction loss has been found to be much less than would be predicted assuming a uniform plane wave aperture distribution and far-field diffraction theory. In fact, machine computations for confocal paraboloidal reflectors show that this loss can be made orders of magnitude less.[2] Fig. 2 shows

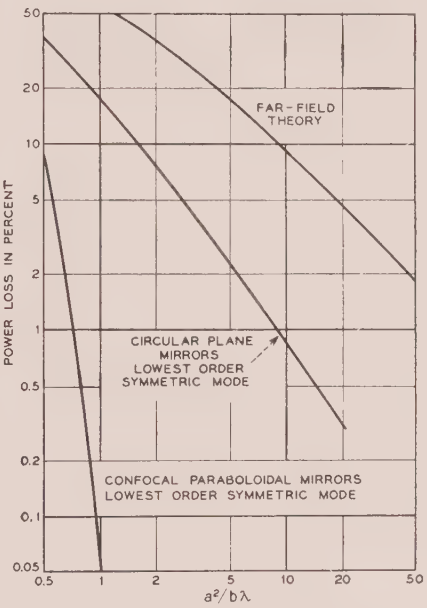

Fig. 2—A comparison of loss per transit for plane and confocal paraboloidal mirrors. A curve based on far-field diffraction theory for circular mirrors with uniform aperture distribution is also shown.

how the loss varies as a function of the parameter $a^2/b\lambda$ for both the plane and confocal paraboloidal mirrors. It now appears likely that the limiting losses in an optical maser will be the reflection losses of the mirrors.

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[2] G. Goubau and F. Schwering, "On the guided propagation of electromagnetic wave beams," presented at URSI-IRE Spring Meeting, Washington, D. C.; May, 1960. G. D. Boyd in a private communication has presented results of his analysis on confocal paraboloidal mirrors of square cross section. Data from these two sources and our own data are in good agreement.

P-N-P Variable Capacitance Diode Theory*

Experimental work on *p-n-p* variable capacitance diodes, in which the center *n* region has no electrical connection [Fig. 1(a)] was recently reported.[1] Steady-state RF operation of such devices as in Fig. 1(b) is analyzed below.[2]

As indicated in Fig. 1(c), the device may be represented by a pair of nearly ideal diodes (low leakage) connected in opposing senses with each shunted by its own barrier capacitance. The interesting part of the operating cycle is at the voltage peaks. For example, as P_1 comes to a voltage peak, diode D_1 is momentarily forward biased, and minority carriers are injected. Some of these carriers recombine, but after the peak the remainder diffuse to the barrier and return to the region of their origin. In the RF steady-state condition, the charge of the injected minority carriers recombined during a cycle must exactly balance the normal reverse leakage of charge from the center region during the cycle (exclusive, of course, of the previously mentioned recovery of injected minority carriers). Aside from the minority carrier and leakage currents, current will flow during most of the cycle directly through the series combination of capacitors C_1 and C_2. The relative charge amplitudes involved in these processes determine the nature of the capacitance variations during the RF cycle and are analyzed below.

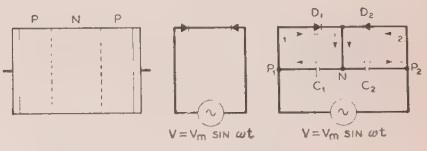

Fig. 1.

The change in charge on the individual junction from minimum to maximum voltage is approximately

$$Q_R = CV_{\max} \tag{1}$$

(C is the capacitance of one junction at maximum voltage and $V_{\max}$ is the peak value of the RF wave). The loss of charge during an RF cycle through normal reverse leakage is given by

$$Q_L = I_R(\mathrm{Av})T, \tag{2}$$

in which $I_R(\mathrm{Av})$ is the average normal reverse leakage current during the cycle and T is the length of the cycle. For most practical junctions, Q_R becomes $> Q_L$ at frequencies above 10,000 cps. At radio frequencies Q_R is therefore much greater than Q_L. Reverse leakage current then affects only the peak stored minority carrier charge.

* Received by the IRE, May 24, 1960; revised manuscript received, July 18, 1960.

[1] J. F. Gibbons and G. L. Pearson, "*P-N-P* variable-capacitance diodes," PROC. IRE, vol. 48, pp. 253–255; February, 1960.

[2] As in footnote 1, it is assumed that transistor action cannot occur in these units, thus permitting the representation as separate diodes in Fig. 1(b).

The steady-state condition requires that the minority carrier charge recombined during its storage interval be equal to Q_L. If Q_F symbolizes the peak stored minority carrier charge, r the fraction of the RF cycle during which the stored charge is close to its peak value, and τ the minority carrier lifetime, this equality can be expressed by

$$Q_L \simeq Q_F r T/\tau. \tag{3}$$

Combining (2) with (3) gives

$$Q_F = I_R(\mathrm{Av})\tau/r. \tag{4}$$

Eqs. (1) and (4) combine to give the ratio of peak injected minority carrier charge to the capacitive charge associated with change of reverse voltage.

$$\frac{Q_F}{Q_R} = \frac{I_R\,(\mathrm{Av})\,\tau}{rCV_{\max}}. \tag{5}$$

When Q_F/Q_R is $\ll 1$, the approximation of constant charge in the center region during the RF cycle is valid. If Q_F/Q_R approaches or exceeds unity, the center region charge will vary greatly during the cycle with consequent effects on the variation of capacitance with voltage.

Q_F/Q_R can be estimated by taking r as the fraction of the cycle during which the voltage is within 0.026 volt (kT/q) of its peak value. For a silicon diode of 10-mmf capacitance, 10^{-8} amperes reverse current, and 10^{-8} seconds minority carrier lifetime, r is less than 0.075 and Q_F/Q_R is less than 10^{-4} for all peak voltages above 1 volt. The constant center region charge is thus a reasonable approximation. Germanium diodes may, in contrast, be designed to make minority carrier storage significant and, in fact, dominant.

Proceeding to the analysis of capacitance for the structure of Fig. 2(a), the barrier charges Q_1 and Q_2 are related to the barrier thickness x_1 and x_2 by

$$Q_1 = qN_D x_1 \qquad Q_2 = qN_D x_2, \tag{6}$$

in which q is electronic charge and N_D the donor concentration of the center region. The charge of the center region is the sum of Q_1 and Q_2, and is therefore proportional to the sum x_1+x_2. Since the center region charge is constant during an RF cycle, x_1+x_2 is also constant. The capacitance of the two barriers in series is therefore similarly constant during an RF cycle, as indicated by the dashed lines in Fig. 2(b).[3]

Although the center region charge does not change appreciably during an RF cycle, an increase of steady-state peak voltage will momentarily cause additional forward biasing at the peak, thus recombining majority carriers from the center region and increasing the stored charge (Q_1+Q_2), thus decreasing the capacitance. A decrease of peak voltage will result in a slow decay $[\tau \simeq Q_R/I_R(\mathrm{Av})]$ of the stored charge to the equilibrium value for the new peak voltage. The solid line in Fig. 2(b) shows the variation of capacitance with peak voltage. It is similar to that of a single step junction, but is symmetrical with respect to voltage.

This analysis suggests that the step junction p-n-p variable capacitance diode with low reverse current and low lifetime is probably a very poor harmonic generator at RF because its capacitance is nearly constant during the RF cycle.

The graded junction case has also been analyzed with results shown in Fig. 3. As in Fig. 2, the dashed lines of Fig. 3(b) show variation of capacitance during the RF cycle, and the solid line shows the variation of capacitance with peak voltage. In Figs. 2(a) and 3(a) the sum of the two-charge areas, Q_1 plus Q_2, must remain constant within the RF cycle. The implications of this for the sum (x_1+x_2), and therefore for capacitance, can easily be seen.

The graded junction unit should generate harmonics. However, the harmonic content should be relatively small and opposite in phase to that predicted from the variation of capacitance with peak voltage. This result is not inconsistent with that of Fig. 2 of footnote 1 (reproduced here as Fig. 4). The data of Fig. 4 were taken using dc bias and a small signal high-frequency capacitance bridge which gives the solid line envelope of Figs. 3 or 4.[4] The capacitance variation shown in Fig. 4 has also been observed using very low frequency (1 cps or less) sinusoidal alternation of the bias voltage,[4] and a capacitance meter. Because this frequency was very much less than the relaxation frequency (reciprocal to reverse bias decay time) of the junction, the bias voltage followed the low-frequency envelope closely, and the observed capacitance variation was effectively a tracing out of the solid line of Fig. 4.

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Murray Hill, N. J.

[4] G. L. Pearson, private communication.

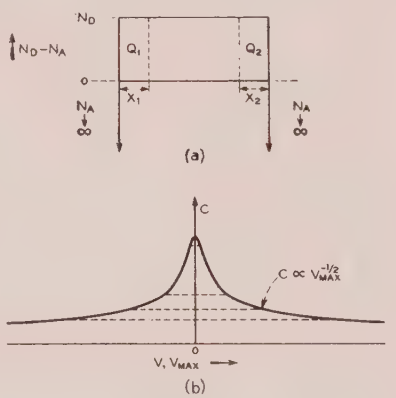

Fig. 2.

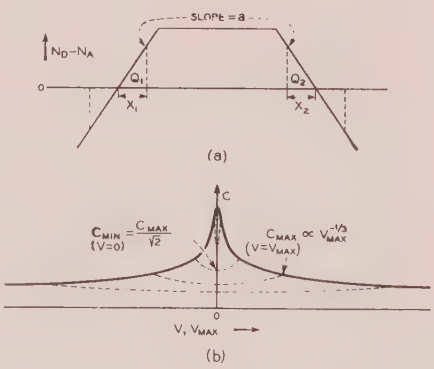

Fig. 3.

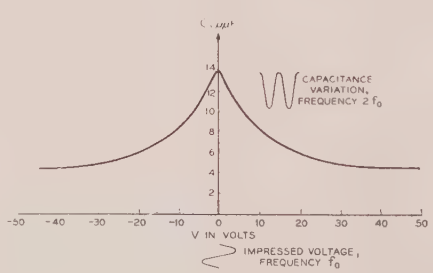

Fig. 4.

Microwave Oscillation and Detection by a Smooth Anode Coaxial Magnetron*

The tunable oscillations and microwave detection properties of a smooth anode, coaxial magnetron have been studied briefly using a tube as shown in Fig. 1. The tungsten filament served as a center conductor for the coaxial line while the outer conductor was the anode. A direct current was passed through the center conductor to heat the tungsten filament. An anode voltage, applied between the two conductors, provided both an axial component (because of the bends in the coaxial line) and a radial component of electric field in the region of the filament. Both field components are necessary for stable oscillation. With one end terminated in a load, the VSWR was below 2 from 1.2–2 kmc.

When the tube was placed in a magnet with the field parallel to the filament, oscillations were observed which were tunable with magnetic field strength. The tuning rate is essentially 2.2 mc per gauss as shown in Fig. 2. There is some deviation from this since change in filament temperature or anode voltage also caused a small frequency shift, the shift being opposite to the change in total space charge. This mode of operation of a magnetron was first investigated many years ago by Okabe, Linder and others[1] using resonant coupling structures. The most recent theoretical work by Pease[2] pre-

* Received by the IRE, July 7, 1960; revised manuscript received July 15, 1960. Supported by the Signal Corps under Contract DA36-039-Sc73188.

[1] K. Okabe, "On the short-wave limit of magnetron oscillations," Proc. IRE, vol. 17, p. 652; April, 1929. E. G. Linder, "Description and characteristics of the end-plate magnetron," Proc. IRE, vol. 24, p. 633; April, 1936.

[2] M. C Pease, "Time-dependent electron flow," *J. Appl. Phys.*, vol. 31, pp. 70–76; January, 1960.

[3] For this device, independent control of center region dc potential through a high RF impedance can give an electronically variable capacitance which is nearly free of distortion for RF voltages smaller than those required for forward biasing or junction breakdown.

dicts an oscillation at $\omega_c/\sqrt{2}$ (where $\omega_c=eB/m$, the free electron cyclotron frequency). This relation agrees with the tuning rate of 2.2 mc/gauss observed experimentally. Further, Pease suggests there may be no start oscillation current condition and we find none; also that the electric field is of constant phase around the filament as we note below. Since this is not a traveling wave interaction and there is no resonant structure, the dc to RF energy exchange is not very efficient. For our particular tube the maximum microwave output was around 100 milliwatts. Attempts to increase this led to a very noisy operation. The power increases about 10 db/ma at the higher power levels until noise generation begins.

A typical tuning run in which the voltage was programmed as the magnetic field was varied to give an approximately constant electron-orbit diameter, gave an output of +8 dbm±4 db from 1–2 kmc. At high magnetic fields, the voltage was maintained at the maximum available value with little change in the output power. Since the voltage tuning rate is approximately 7 mc per 100 volts and the current tuning rate is about 5 mc per ma of plate current, the power supply stability requirements are not severe. The magnetic field would doubt ess provide the main source of frequency instability. With a stable magnetic field the oscillations were as narrow in bandwidth as those obtained from the klystron powered signal generator, even though the spatial variation of magnetic field should lead to a considerably larger bandwidth. The spectrum remained narrow even when the magnetic field was purposely made spatially inhomogeneous.

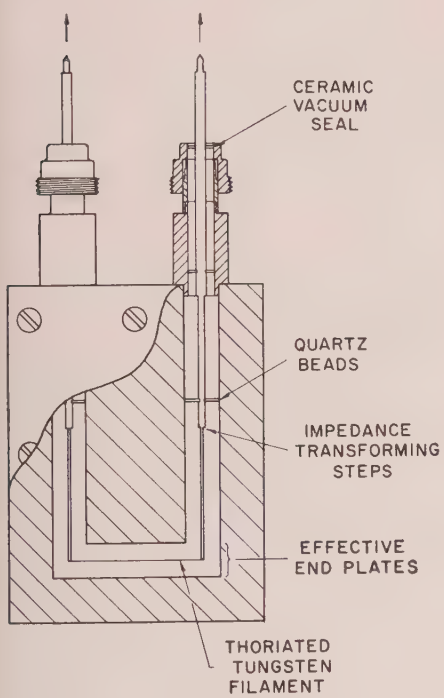

Fig. 1—Coaxial magnetron tube.

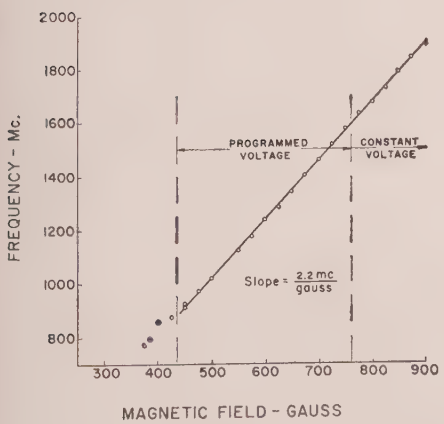

Fig. 2—Coaxial magnetron tuning range.

Spurious oscillations were more than 60 db down except for the second harmonic (−10 to 15 db) and a signal at ω_c. For conditions of maximum power at $\omega_c/\sqrt{2}$ the power at ω_c was down more than 50 db, but for lower power the relative amount of ω_c power increased. In measurements on a tube equipped with probes, it was determined that the electric field of the $\omega_c/\sqrt{2}$ oscillations had no phase variation around the cathode, *i.e.*, the same electric field configuration as a TEM mode. Unfortunately, no phase measurements of the ω_c oscillations were made, but there was other evidence to indicate that the mechanisms of the two oscillations were completely different.

When a signal generator was connected to one end of the tube, the output of the tube and signal generator could be observed simultaneously on a spectrum analyzer connected to the second output. When the signal generator was tuned through the ω_c oscillation no interaction between the two signals was observed even though the transmitted signal from the generator could be made 20 to 40 db stronger than the signal from the tube. On the other hand, when the generator frequency approached the $\omega_c/\sqrt{2}$ oscillation very strong coupling between the two signals was observed. For generator power equal to or greater than the oscillation output, the tube oscillation was captured when the generator frequency came within 1 to 5 mc of the tube frequency. As the generator was tuned only one output was observed until the oscillator snapped back to its original frequency as the generator frequency moved away. When the generator power was much smaller than the oscillator power, side bands were formed as the generator frequency approached $\omega_c/\sqrt{2}$ This effect could be seen even when the generator power was very small.

The interaction of an external signal with the oscillations led us to look for a change in anode current as a signal generator was tuned through the oscillation frequency. As expected, a change in current occurred but we were quite unprepared for the sensitivity obtained. The presence of a signal was observed by noting the potential change across a 10^6 ohm resistor in the anode supply line. The noise inherent in our power supply was quite large, but the presence of signals of −65 to −70 dbm could be readily observed. Detection sensitivity was best for a low power oscillation condition. Under one particular set of conditions a pulsed signal of −87 dbm was detected. This compares favorably with the cyclotron detector mentioned by Turner.[3]

When a 30-mc gain set was connected to the tube, a signal was observed when the generator frequency was ±30 mc from the oscillator. The minimum detectable signal level by this technique was about −45 dbm, probably because the tube and associated apparatus presented a very bad match at 30 mc.

Fig. 3 shows a polarity and amplitude of the detected response as a function of frequency around the oscillation point. The

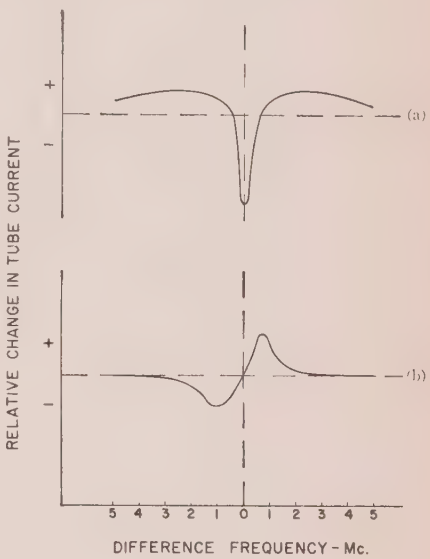

Fig. 3—Tube current variation with coincidence of external signal frequency and oscillation frequency.

difference in the pattern is probably attributable to the distinctly different reactions of the oscillator, mentioned previously, which depend on whether the detected signal is significantly larger or smaller than the amplitude of the oscillations. Fig. 3(a) shows the response for oscillator power greater than signal power, while Fig. 3(b) is for the opposite case. Finally, it was found that the change in anode current is quite accurately proportional to the square root of the signal power or directly proportional to the electric field strength of the detected signal. It is felt that a significant improvement in detection sensitivity could be made through the use of an improved power supply.

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[3] F. M. Turner, "A broad-band cyclotron resonance RF detector tube," Proc. IRE, vol. 48, p. 890; May, 1960.

New Microwave Tube Devices "Fawshmotron" Using the Fast Electron Wave*

Parametric amplification which utilizes an electron beam as a nonlinear element is classified into two general divisions. One is the parametric amplifier using the fast wave in a straightforward electron beam proposed by Ashkin *et al.*,[1] which is called the longitudinal beam parametric amplifier. The other is the quadrupole amplifier developed by Adler *et al.*[2,3] and it is called the transverse beam parametric amplifier.

The method described here belongs to the latter, which uses the fast electron wave that appears when electrons travel in the axial direction with transversal simple harmonic motion. The fast electron wave of this type is not only used as the parametric amplifier, but also is used as a microwave oscillator or amplifier without slow wave circuits. We call such a microwave tube operated by the fast wave produced by simple harmonic motion of electrons the "Fawshmotron," taking the first letters of "Fast Wave Simple Harmonic Motion."

Fig. 1 shows an example of electrode construction in which a potential distribution is a parabola, and this potential distribution is expressed as

$$V(x, y) = V_a\left\{\left(\frac{x}{l}\right)^2 - \left(\frac{y}{l}\right)^2\right\}. \qquad (1)$$

When the electrons travel in the z direction, describing a simple harmonic motion in the y direction within the space of this potential distribution, a pair of electron waves are produced. These waves are called the fast and the slow electron wave and their phase velocities are

$$v_p = \frac{u_z}{1 \mp (\omega_0/\omega)}, \qquad (2)$$

$$\omega_0^2 = 2\eta V_a/l^2, \qquad (3)$$

where

u_z = velocity of electron,
ω = signal angular frequency,
η = charge to mass ratio of electron,
l = distance between electrode A and the origin,
V_a = potential of electrode A.

In (2), the upper sign is applied in the case of fast wave and the lower sign in the case of slow wave. It should be noted that the phase velocity of the fast wave may be made even infinite or negative, as the three conditions $\omega_0/\omega \lessgtr 1$ can readily be realized. Fig. 2 shows the relation between the phase velocities of electron waves and the values of ω_0/ω. We can construct a parametric amplifier which employs the above mentioned fast wave just like the Adler type.

The general construction of this type is shown in Fig. 3. In this device, the potential of electrode A is so chosen that ω_0 is equal to the signal angular frequency ω. The signal applied to the input circuit causes the electrons to describe simple harmonic motions. They reach a maximum amplitude at the end of the input circuit. At this point, the entire signal power supplied to the input circuit is converted into the kinetic energy of electrons.

The operation of the output circuit and the mechanism of noise reduction are quite similar to those of the quadrupole amplifier. Parametric amplification of the simple harmonic motion of electrons can be realized by applying the pumping signal between electrode A and electrodes O and O' which are commonly connected to each other. For convenience, this type of parametric amplifier is called the "A-type Fawshmotron."

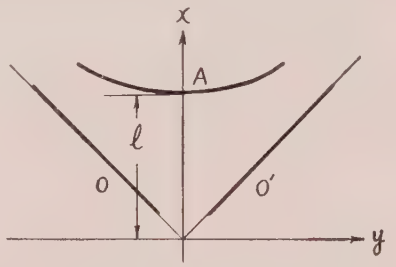

Fig. 1—An example of the electrode structure which gives a parabolic potential distribution.

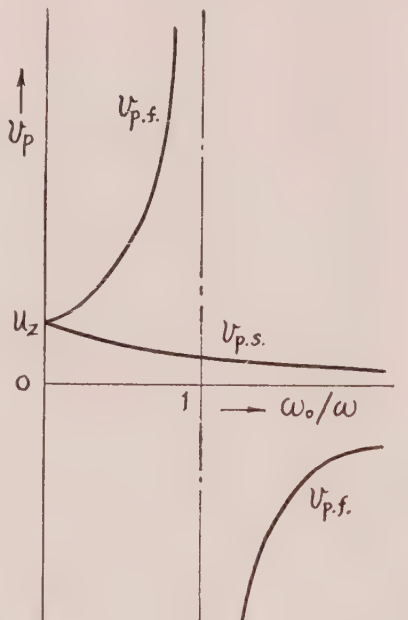

Fig. 2—Phase velocities of the fast and the slow waves as a function of ω_0/ω.

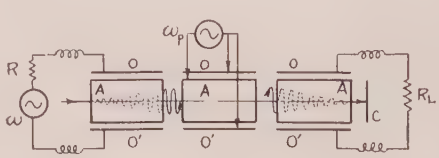

Fig. 3—Parametric amplifier (A-type Fawshmotron).

As shown in Fig. 2, the fast wave of this type may have infinite phase velocity when $\omega_0/\omega = 1$, and it may be easily synchronized with the circuit wave propagating along a circuit having no slow wave structure.

In the region of $\omega_0/\omega < 1$, the phase velocity of the fast wave is positive. It is synchronized with that of the circuit wave, whereby the ac power-flow of the fast wave and of the circuit wave are in the same direction. If two lossless modes with ac power-flow in the same direction are coupled together, the ac power can flow back and forth between them, as is shown by the theory of the coupling of two modes of propagation. By coupling these modes, therefore, microwave oscillator or amplifier can be obtained. We call this type of tube the "B-type Fawshmotron." This type of microwave tube corresponds to the "traveling wave Peniotron" proposed by Koike[4] or the "no teeth magnetron" developed by Morita,[5] which utilize the fast wave of cyclotron wave.

In Fig. 2 it is found that the fast wave in the region of $\omega_0/\omega > 1$ has negative phase velocity. Therefore, the power flow of the fast wave and that of the circuit wave are in opposite directions, when the phase velocities of these waves are synchronized together. By utilizing the coupling between these modes the backward wave oscillator is obtained, and this is called the "C-type Fawshmotron." In microwave tubes which utilize the cyclotron wave, the backward wave oscillator proposed by Pantell[6] corresponds to this case.

The case of $\omega_0/\omega = 1$ is the Osaka-tube operating with the fundamental mode, which was studied by Okabe,[7] and the Strophotron.[8] In the cyclotron wave, the "cyclo-magnetron oscillator"[9,10] corresponds to this case.

The author wishes to express his sincere thanks to Prof. S. Sonoda for his interest and encouragement.

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* Received by the IRE, July 11, 1960.

[1] A. Ashkin, T. J. Bridges, W. H. Louisell and C. F. Quate, "Parametric electron beam amplifiers," 1958 IRE WESCON Convention Record, pt. 3, pp. 13–22.

[2] R. Adler, "Parametric amplification of the fast electron wave," Proc. IRE, vol. 46, pp. 1300–1301; June, 1958.

[3] R. Adler, G. Hrbek and G. Wade, "The quadrupole amplifier, a low-noise parametric device," Proc. IRE, vol. 47, pp. 1713–1723; October, 1959.

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[5] K. Morita and Y. Suematsu, "Super-wide-band magnetron oscillator," presented at the monthly meeting of the Professional Group on Microwave Tubes in Japan; February 17, 1959.

[6] R. H. Pantell, "Backward-wave oscillations in an unloaded waveguide," Proc. IRE, vol. 47, p. 1146; June, 1959.

[7] K. Okabe, M. Hisida, and K. Owaki, "A new method of microwave oscillations," *J. Inst. Comm. Engrs., Japan*, p. 850; July, 1936.

[8] H. Alfven and D. Romell, "A new electron tube, the strophotron," Proc. IRE, vol. 42, pp. 1239–1242; August, 1954.

[9] Y. Matsuo, K. Ebisutani, and Y. Cho, "Microwave oscillator," presented at the Spring Meeting of the IEE in Japan, May, 1958.

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Negative-Resistance Distributed Amplifier*

Two well-known types of distributed amplifiers are those that employ voltage-controlled current generators[1] for their active elements and those that employ nonlinear reactances.[2] In addition to the two types cited, realizable negative resistance may be employed in an iterative low-pass network in order to obtain wide-band amplification. It is the purpose of this letter to describe the theory of such a negative-resistance distributed amplifier.

For two-terminal pair networks, the following theorems have been derived.[3]

Theorem I: In a lossless (passive), linear, two-terminal pair network that is terminated in a positive or negative resistance R_2, the input impedance Z_1 of the remaining pair of terminals will possess the following property in relation to the sign of R_2:

$$Z_1 = R_a + jX_a \qquad \text{for } R_2 = R_b$$
$$Z_1 = -R_a + jX_a \qquad \text{for } R_2 = -R_b.$$

Theorem II: In a lossless (passive), linear, two-terminal pair, bilateral network that is terminated in a positive or negative resistance R_2, the voltage transfer ratio E_2/E_1 will possess the following property in relation to the sign of R_2:

$$E_2/E_1 = G \qquad \text{for } R_2 = R_b$$
$$E_2/E_1 = G^* \qquad \text{for } R_2 = -R_b,$$

where G^* is the complex conjugate of G.

Theorem I, in particular, may be applied to the well-known π or T constant-k low-pass filter sections. Below cutoff, constant-k low-pass filters have a characteristic impedance R_0 that is purely resistive, and Theorem I reveals that this characteristic resistance can either be positive or negative. R_0 can also be shown to be either positive or negative if we consider that a double-valued function is obtained in the derivation of R_0 by well-known means.

Because the characteristic resistance of a π or T constant-k low-pass filter can be either positive or negative, it is possible to devise a distributed amplifier that employs negative resistance for its active elements and that has an image termination associated with every filter section employed. The configuration for the distributed amplifier and a first-order theory of operation will be presented. In order to simplify the presentation, a distributed amplifier that employs four negative resistances will be described; it should be kept in mind, though, that the method of employing the negative resistances in the amplifier can be extended to any number of negative resistances. Furthermore, the filter sections need not be of the π or T constant-k types since the concept presented applies equally well to other networks that are made up of iterative sections.

Since it is impossible to realize a negative resistance at high frequencies without a significant shunt capacitance, a negative resistance device consisting of a resistor shunted by a capacitor will be assumed.

In order to simplify the explanation of how the distributed amplifier functions, the distributed amplifier will be broken down into a number of smaller networks. The distributed amplifier consists of a *main network* and a number of *auxiliary networks*. There are as many auxiliary networks as there are negative resistances in the main network.

Fig. 1 shows the main and the four auxiliary networks that comprise the distributed amplifier. The main network consists of seven π-type low-pass constant-k filter sections joined together. At each odd-numbered node there is a negative resistance, and at each even-numbered node there is indicated the positive resistance that must be added in order to provide an image termination for each section. The maintenance of an image termination for each π section will make the input impedance of the main network equal to R_0. R_0 is the positive characteristic resistance of the sections that comprise the main network. Note the following carefully:

1) The π section between nodes 1 and 2 (called π section 1-2) is terminated in its negative characteristic resistance, $-R_0$.
2) When $R_0/2$ is added at node 2, it is apparent that π section 2-3 is terminated in its positive characteristic resistance, R_0.
3) π section 3-4 is terminated in its negative characteristic resistance, $-R_0$.
4) The main network may be continued indefinitely in the manner just described. It is apparent that all the π sections will be terminated in their characteristic resistance, whether it be *positive* or *negative*.

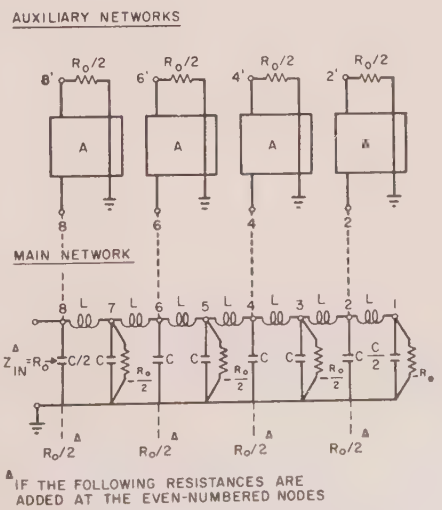

Fig. 1—Negative-resistance distributed amplifier.

The resistances, $R_0/2$, that must be added to each even-numbered node of the main network, might simply be resistors that are shunted across the even-numbered nodes, but this would not be at all practical. The power output of the amplifier consists of the power delivered to each resistance that is connected to an even-numbered node. If at all possible, it would be advisable to sum all of the output powers available at the even-numbered nodes in order to obtain a single output.

For the moment, consider the main network to have resistances $R_0/2$ added to each even-numbered node. Every π section is now terminated in its characteristic resistance whether it is positive or negative, and each π section has a magnitude of one for its voltage transfer ratio; this must be so because of energy conservation in the lossless (passive) π sections. Thus, the *magnitude* of the voltage at every node of the main network is the same as the magnitude of the input voltage to the main network. Furthermore, if Theorem II is considered, it is apparent that the voltages at the *even-numbered* nodes are equal not only in magnitude but also in *phase;* the same thing is true for all the *odd-numbered* nodes.

Assuming that resistances $R_0/2$ are still connected to the even-numbered nodes, it is possible to join all the even-numbered nodes without any interaction. The total output load for the amplifier is then $R_0/(2N)$, where N is the number of negative resistances employed in the amplifier.

For an amplifier that operates in the VHF to microwave frequency region, each of the even-numbered nodes on the main network must be connected to auxiliary networks A having the following ideal properties:

a) characteristic impedance $= R_0/2$,

b) lossless (passive), and

c) a phase shift that varies directly with frequency.

All the auxiliary networks are identical, and each one of the auxiliary networks is terminated in its positive characteristic resistance. If the main network is now considered with the properly terminated auxiliary networks joined to the even-numbered nodes, it is apparent that the voltages at the output terminals of all the auxiliary networks are identical. Since this is so, the primed even-numbered nodes may all be joined together without any interaction.

The generalized equivalent circuit for a distributed amplifier that employs N negative resistances is shown in Fig. 2 (next page). It is obtained by joining together all nodes having identical voltages in the amplifier. From the equivalent circuit the following is evident: voltage gain $= 1$, power gain $= 2N$, input resistance $= R_0$, and output load resistance $= R_0/2N$.

In practice, the amplifier will be composed of tunnel diodes and either strip-line or coaxial transmission sections. Operation from dc into the microwave frequency region with power gains of at least ten should eventually be possible.

* Received by the IRE, December 21, 1959; revised, January 14, 1960.

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[2] P. K. Tien and H. Suhl, "A ferromagnetic traveling wave amplifier," PROC. IRE, vol. 46, pp. 700–706; April, 1958.

[3] C. A. Skalski, "Negative-Resistance Distributed Amplifier," Norden Labs., White Plains, N. Y., Rept. D4 R 0005; September 30, 1959.

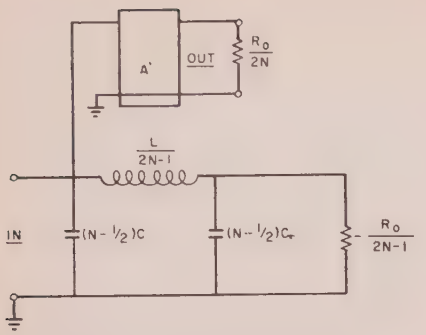

Fig. 2—Equivalent circuit for a negative-resistance distributed amplifier that employs N negative resistances. A' consists of N auxiliary networks in parallel.

The author wishes to thank L. Chanzit for his encouragement of this work and Dr. L. Kosowsky for a number of stimulating discussions concerning the theory.

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Traffic Efficiencies in Congested Band Radio Systems*

In a recent paper[1] an analysis of congested band radio operation, such as is often encountered in amateur and military situations, was presented. Statistical methods were used to show that broad-band techniques are in a certain sense superior to presently used narrow-band methods. It is the purpose of this note to show that a somewhat different point of view will find the narrow-band system always superior, although a similar statistical approach is used and all assumptions remain unchanged.

In essence, the argument is as follows: Costas[1] indicates that wide-band systems are more reliable than narrow-band for some ranges of band loading, *providing the desired signal is at least of average power at the receiver*. It will be shown here that if it is desired to communicate with a particular individual *regardless of his received signal power* (which is the same as selecting a transmitting station at random), the probability of receiving his transmission is always higher if narrower-band systems are used (see Fig. 1).

Three situations are considered.[1] The first is exclusive use by all stations of single-sideband (SSB) suppressed-carrier transmission, the second is exclusive use of double-sideband (DSB) suppressed-carrier transmission, and the third is a very wide-band frequency diversity system. The following assumptions were made, and will be retained in this analysis:

1) A signal is said to be received if the signal-to-noise ratio at the post-detector is at least 0 db.
2) Received signal power densities are distributed according to the chi-square distribution, *i.e.*,

$P_1(S) = \text{Prob}$ [Power received from one station $\leq S$]

$$= \int_0^S \frac{e^{-x/2}}{\sqrt{2\pi x}}\, dx$$

and therefore

$P_n(S) = \text{Prob}$ [Total power received from n stations $\leq S$]

$$= \int_0^S \frac{x^{(n/2-1)}e^{-x/2}}{2^{n/2}\Gamma(n/2)}\, dx.$$

3) Signals are placed at random within the spectrum (for SSB and DSB) and the average number in a bandwidth interval B is kB, where k is the average number of signals per unit bandwidth. The probability that a bandwidth B' contains n signals is then given by the Poisson distribution

$$\pi(n, B') = \frac{(kB')^n}{n!} e^{-kB'}.$$

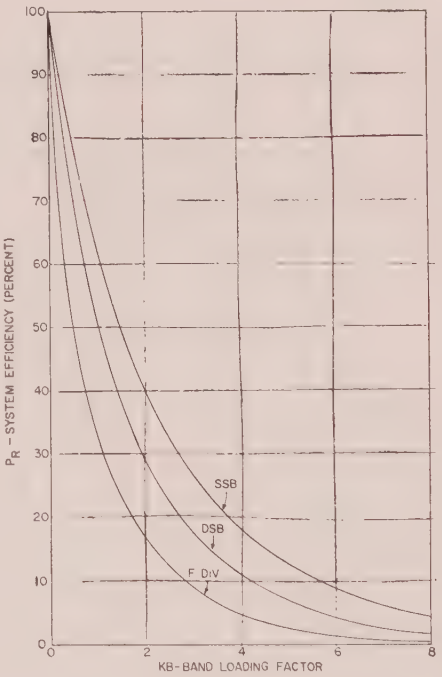

Fig. 1—System efficiency vs band loading.

Now suppose we wish to communicate with a particular individual. The power of his signal at our receiver is not known, and therefore can only be described by the distribution $P_1(S)$. This signal must compete with n unwanted signals within his frequency band which have a total power given by $P_n(S)$. The probability that the ratio of the wanted to unwanted signals (S/N) is sufficiently large to enable the wanted signal to be received must now be found.

Let ξ_1 be the random variable whose distribution function is given by $P_1(S)$, and ξ_n be the random variable whose distribution function is given by $P_n(S)$. Then we require

$$\text{Prob}\left[\xi_1 \geq \frac{S}{N}\,\xi_n\right] = \text{Prob}\left[\frac{\xi_1}{\xi_n} \geq \frac{S}{N}\right]$$

for each n, where n is the number of interfering signals. If this is then summed over all n, where the probability of n interfering signals (for SSB and DSB) is given by $\pi(n, B')$, the probability that the ratio of wanted to unwanted signal is at least S/N is obtained. The results in the three cases are as follows:

1) SSB—In this case, S/N must be at least 1, and the bandwidth is B. Then

Probability of signal reception =

$$P_R = \sum_{n=0}^{\infty} \text{Prob}\left[\frac{\xi_1}{\xi_n} \geq 1\right]\pi(n, B),$$

where

$$\text{Prob}\left[\frac{\xi_1}{\xi_n} \geq x\right]$$

is a tabulated distribution.[2]

2) DSB—In this case S/N must be at least 1/2, since there is a 3-db gain due to coherent adding of the two sidebands, while the interfering bandwidth is $2B$. Accordingly,

$$P_R = \sum_{n=0}^{\infty} \text{Prob}\left[\frac{\xi_1}{\xi_n} \geq \frac{1}{2}\right]\pi(n, 2B).$$

3) Frequency diversity—Here the sum of the interfering signals is constant with intensity kB, while the wanted signal is distributed according to $P_1(S)$. Then

$$P_R = \text{Prob}\left[\frac{\xi_1}{kB} \geq 1\right]$$

$$= \text{Prob}\left[\xi_1 \geq kB\right] = 1 - P_1(kB).$$

P_R has been calculated for each of the three cases and curves of P_R vs kB are given in Fig. 1. Examination of this figure shows that P_R is always highest for the narrowest bandwidth situation (SSB) and gets progressively lower as the bandwidth is increased. This result can be interpreted in several ways. One reasonable interpretation might be that the proportion of stations which can be received is greatest if narrow-band techniques are used. Another is as follows: If a certain number of messages, N, are transmitted during a period of time, T, by all transmitters, then the number of such messages which can be received is $P_R N$, providing enough receivers are available to accept them. If N/T is considered to be the offered traffic load, then $P_R N/T$ will be the carried traffic, and

$$P_R = \frac{\text{carried traffic}}{\text{offered traffic}}$$

is a reasonable measure of the efficiency of the system from a traffic standpoint. If only those messages which are of interest to a particular receiver are included in N, the same argument yields an identical result.

If P_R is accepted as a measure of the system efficiency, then Fig. 1 shows that

* Received by the IRE, March 21, 1960.

[1] J. P. Costas, "Poisson, Shannon, and the radio amateur," Proc. IRE, vol. 47, pp. 2058–2068; December, 1959.

[2] H. Cramer, "Mathematical Methods of Statistics," Princeton University Press, Princton, N. J., ch. 18; 1946.

narrow-band techniques are clearly superior at all values of band loading.

The same analysis was carried out using a negative exponential distribution for $P_1(S)$, and no significant differences in the results were obtained.

It is to be emphasized that nothing in this note contradicts the results of Costas.[1] The viewpoint is merely changed from that of the individual receiving a signal of some particular magnitude to that of the individual receiving a signal from a particular transmitter. It has been shown, however, that narrow-band systems can in some sense be thought superior even if interference is taken for granted and a statistical approach is used, which result was nowhere suggested by Costas.

In conclusion, it would appear that in congested band radio systems, if the primary objective is to receive a few strong stations reliably, a wide-band system may prove to be superior in certain cases. If, however, it is desired to communicate with particular individuals without reference to their received powers, a narrow-band system will provide a higher over-all system efficiency.

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Author's Comment[3]

It appears from Weber's second and concluding paragraphs that I have failed to make clear what I believe to be the true significance of the circuit reliability curves shown in Part III of the original paper. Voice operation in a congested band was considered and circuit reliability was computed for three different modulation methods as a function of band loading and for different assumed values of desired signal strength. The curves (Figs. 2–5) show that the three modulation methods give *different* results. I attach no particular importance to the fact that under some conditions DSB and the broad-band system show a higher circuit reliability than SSB. However, these curves do indicate rather clearly that for congested-band operation:

1) The world will not come to an end if transmission bandwidths in excess of those corresponding to SSB are employed.
2) If the congestion becomes heavy and if the desired signal strength is low, neither the broad-band nor the narrow-band system will be satisfactory. The first system eliminates the weak signal completely. The second system permits reception only for a small percentage of the time (during rare and unpredictable lulls in the interference level in the channel) so that very little if any message handling will be possible.
3) If messages *must* be handled from stations with low signal levels, the data rate will have to be reduced; voice cannot be used.

[3] Received by the IRE, April 15, 1960.

The above-mentioned curves together with some additional discussion in the paper allow one further point to be made:

4) In military communications where traffic must be handled with *reliability* under various conditions of band loading, jamming, and desired signal level, the following must be done:
   a) The user must be willing to reduce data rate to values commensurate with the band loading, jamming, and desired signal level parameters. In many cases this eliminates voice completely.
   b) Broad-band techniques must be used.

In answer to Weber, then, no claim was made that broad-band systems are better than narrow-band systems in so far as Figs. 2–5 in the paper and points 1–3 above are concerned. With regard to point 4, broad-band systems are not only better but they are the only answer if any value at all is attached to *reliability*. This evaluation assumes that the military user is willing to lower his data rate requirements when necessary. If he is not so willing, then I believe he is in for serious trouble regardless of whether narrow- or broad-band techniques are used.

Weber's comment that only desired signal strengths above mean level were considered in the paper can easily be explained. It is obvious from the circuit reliability curves presented that little would be gained by showing the results for desired signal levels below the average. At heavy band loading values (say $kB>4$) SSB shows a circuit reliability of less than 20 per cent for average strength signals and, of course, the reliability will be even poorer for desired signals of lesser strength. Thus there seemed to be little point in presenting additional curves. When a station can only be heard 10 or 20 per cent of the time at unpredictable moments, you aren't going to do much business with this fellow, especially if your transmissions are equivalently weak in his receiver. The circuit reliability may not be zero (as it would be for broad-band operation under the same conditions), but for all practical purposes it may as well be.

Weber has accepted the results in the paper but he proposes a performance measure P_R which he has called the "system efficiency." He shows that if such a measure is used, narrow-band systems will always be superior. Where I have shown probability of reception (circuit reliability) for different desired signal levels, Weber obtains an overall *average* circuit reliability, P_R, by considering all possible desired signal strengths and their relative frequency of occurrence. One physical interpretation of P_R would be as follows: A large number of signals is selected at random from the congested band. Each signal is monitored for, say, 15 minutes of actual transmission time and the total time that reception was possible is noted in each case. The quantity P_R would then be obtained by taking the ratio of the total reception time logged to the total transmission time monitored. With such a measure, narrow-band systems should show an advantage. In the random selection of signals many more weak ones will be chosen than strong ones. Now the weak signals will perform poorly but each will contribute something to P_R in narrow-band operation. With broad-band operation these weak signals will not be heard at all and will contribute nothing to P_R. Whether or not it makes sense to credit the weak signal in the narrow-band case with the reception of isolated words or sentences is a matter of opinion. It is on this point that I question Weber's P_R as a measure of performance.

Weber gives two interpretations of P_R. The first of these is that P_R represents the proportion of all stations that can be received. To this I would add "at any time." Thus if a search across the congested band is made and all stations which can be received are counted, one will naturally find the score higher with narrow-band usage. Many weak signals will be received in narrow-band operation that will not be found in the broad-band case. However, consider a second sampling in the narrow-band case after, say, a one-minute interval has elapsed. The proportion of stations received will still be about the same but, as far as the weak signals are concerned, the particular ones received in the second sampling will generally be different from those of the first. Thus we are not so much concerned with how many stations can be received; it is more important to know what to expect from a particular station from whom we must take traffic. For this reason I feel the curves in the paper are of somewhat more significance to the user than the P_R presentation. The second interpretation given by Weber is that P_R represents the ratio of carried traffic to offered traffic. Here again the weak signals in narrow-band operation are being given "part score" for what will most likely be isolated parts of messages.

The P_R criterion presented by Weber is an interesting one and if accepted would result in a preference for narrow-band over broad-band operation. For the reasons cited above I do not accept this measure of performance but this is a matter of personal opinion and judgment; it is not a question of right or wrong.

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The Backfire Antenna*

Recently a new type of a directional antenna has been described.[1] This antenna consists of a conventional Yagi array and a plane reflector placed at the forward edge perpendicularly to the axis of the array.

The reflector causes the main beam to point in the backward direction, as contrasted with the forward beam of a regular Yagi array. Also, and this is the main ad-

* Received by the IRE, April 12, 1960.

[1] H. W. Ehrenspeck, "The backfire antenna, a new type of directional line source," Proc. IRE, vol. 48, pp. 109–110; January, 1960.

vantage of the antenna, there is a considerable increase in directivity compared with a conventional Yagi array of the same length. The explanation for this increase in directivity is based on the Hansen-Woodyard condition and a resulting linear relationship between the gain and the length of a linear array. It is claimed that, since the reflector effectively doubles the length of the array, a gain increase of 3 db should be expected. Experiments are described which verify this principle qualitatively, but, in general, gain increases are obtained which are well in excess of the predicted value of 3 db.

This note is concerned with a criticism of the interpretation of the Hansen-Woodyard condition as applied to parasitic arrays. As a result, a different explanation is offered of the gain increase effect observed on the backfire antenna.

Strictly speaking, the Hansen-Woodyard condition, as originally derived,[2] applies to a linear array of equispaced *radiating* elements, each of which is excited by a current proportional to $\exp(jkz)$ where z is the distance along the axis of the array, and k is a constant to be chosen for maximum directivity. This type of distance dependence is about the only thing which excited and parasitic arrays may have in common, which is proven for the latter case by Floquet's theorem. Arrays of excited elements, that are long in terms of wavelengths, exhibit a linear relationship between gain and length which is given by Hansen and Woodyard by

$$\text{gain} = 4.83\frac{2Z_0}{\lambda},$$

where $2Z_0$ is the length of the array. This relationship is derived from the vector summation of a large number of radiating sources and cannot be applied to short parasitic arrays.

The basic difference between an excited and a parasitic array is that in the former each element is an active, radiating element, while in the latter each element is a passive, nonradiating storage element. An array of such elements forms a surface wave transmission line which carries energy from the exciting element to the radiating aperture formed by the abrupt termination of the line. The radiation pattern of such a parasitic array is determined by the field distribution across the aperture and by the direct radiation from the feed.

For short arrays, however, another effect must be accounted for. Besides the dominant dipole mode, radiation modes are excited in the vicinity of the feed, and a substantial length of line is required before a dominant mode surface wave of high purity can be established. If the line is terminated before this situation is reached, the actual radiation pattern shows considerable deviation from the pattern calculated on the assumption that only the dominant mode exists. This is believed to be the reason for the observed effect that the gain of *short* parasitic arrays increases with length and converges on the calculated value for arrays some 8 λ long.[3,4]

In the case treated by Ehrenspeck, a 2 λ long Yagi antenna was used and its effective length increased to 4 λ by the use of a reflector; therefore, the increase in gain can be taken as resulting from the increase in the purity of the surface wave mode. In order to test this hypothesis, longer arrays may be employed and it is expected that, for a certain value of array length and above, very little gain increase will be noted when the array is turned into a backfire antenna. This value will depend, among others, on the efficiency of the exciting element.

It should be realized that what has been said does *not* imply that the Hansen-Woodyard condition cannot be applied to parasitic arrays. The argument has dealt exclusively with the application of the linear gain-length relationship, and not with the original condition.

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[3] R. M. Fishenden and E. R. Wiblin, "Design of Yagi aerials," *Proc. IEE*, vol. 96, pt. III, pp. 5–12; January, 1949.
[4] J. O. Spector, "An investigation of periodic rod structures for Yagi aerials," *Proc. IEE*, vol. 105, pt. B, pp. 38–44; January, 1958.

A Note on a Pattern Recognition Scheme*

In a recent article,[1] a pattern recognition scheme is described in which the crucial requirement is the unique association of each pattern with a set of two-dimensional properties. The purpose of this note is to present bounds to the number of properties needed under such a scheme.

Consider a finite set of patterns, each having a finite number of distinct forms, or "variants." Also consider a set of properties, or "features," such that each variant is associated with a subset of these features. For the patterns to be recognizable, the subset of features associated with any variant of one pattern should differ from the subset associated with any variant of any other pattern. The smallest set of features which fulfills this requirement will be called "minimal."

Theorem. Let M be a set of m patterns, the i th pattern having v_i variants. Let N be a minimal set of n features associated with M. Then

$$\log_2 m \leq n \leq q - 1$$

where

$$q = \sum_{i=1}^{m} v_i.$$

Proof. Designate the variants of the first pattern by $r_1, \cdots, r_{v_1}$, the variants of the second pattern by $r_{v_1+1}, \cdots, r_{v_1+v_2}$, and so on. Construct a matrix $[\delta_{ij}]_{q\times n}$, where

$$\delta_{ij} = \begin{cases} 1 & \text{if variant } r_i \text{ possesses feature } j \\ 0 & \text{otherwise.} \end{cases}$$

Partition the rows of this matrix into m subsets, each representing all the variants of a different pattern. If the patterns are recognizable, no two rows appearing in two different subsets can be identical. Merging every group of identical rows into one results in a $q'\times n$ matrix, where

$$m \leq q' \leq q. \tag{1}$$

The resulting matrix is a "pattern matrix," as defined by Gill,[2] for which, according to this reference, the following is true:

$$\log_2 q' \leq n \leq q' - 1. \tag{2}$$

Combining (1) and (2) yields the stated theorem.

Cases where the lower bound is achieved with equality can be readily constructed for every m which is a power of 2. Matrix (3) shows that the upper bound is achievable with equality for any q, regardless of the v_i values. This can be verified by observing that the deletion of any single column in the matrix causes two rows, representing variants of two different patterns, to become identical.

$$\begin{array}{l} \\ v_1 + \cdots + v_{m-1} \\ \\ \\ \\ \\ \\ \\ \\ \\ v_m \\ \\ \\ \end{array} \overbrace{\begin{bmatrix} 100\cdots0000\cdots00 \\ 010\cdots0000\cdots00 \\ 001\cdots0000\cdots00 \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ 000\cdots1000\cdots00 \\ 100\cdots0100\cdots00 \\ 100\cdots0010\cdots00 \\ 100\cdots0001\cdots00 \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ 100\cdots0000\cdots10 \\ 100\cdots0000\cdots01 \\ 000\cdots0000\cdots00 \end{bmatrix}}^{q-1} = \tag{3}$$

The designer's obvious objective is to make m as close to the lower bound as possible, and hence to find a set of features such that a) all variants of the same pattern possess the same subset of features, b) differing patterns possess different subsets of features, and c) the number of features in which any two dissimilar patterns differ is as small as possible (but greater than zero). To the extent that these requirements are incompatible for any given set of patterns, n will be closer to the upper bound in the stated theorem.

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[2] W. W. Hansen and J. R. Woodyard, "A new principle in directional antenna design," PROC. IRE, vol. 26, pp. 333–345; March, 1938.

* Received by the IRE, December 15, 1959.
[1] S. H. Unger, "Pattern detection and recognition," PROC. IRE, vol. 47, pp. 1737–1752; October, 1959.

[2] A. Gill, "Minimum-scan pattern recognition," IRE TRANS. ON INFORMATION THEORY, vol. IT-5, pp. 52–58; June, 1959.

Evidence of Satellite-Induced Ionization Effects Between Hemispheres*

It has been found at a remote CW station at the Ohio State University Radio Observatory, Columbus, that large signal enhancements often occur at times related to the near approach of an artificial earth satellite.[1–3] The magnitudes of the enhancements indicate reflection cross sections which are usually at least several orders of magnitude greater than either the physical cross section of the satellite or the effective aperture of its antenna. The implication is that large amounts of ionization are induced by the satellite. In these observations it has been convenient to monitor station WWV near Washington, D. C., at a distance of 330 miles from Columbus.

The ionization effect is a sporadic one and seems to be dependent upon several factors. One of these appears to be the position of the satellite with respect to the auroral zones and the Van Allen belts. The purpose of this note is to present some evidence which suggests that some CW-reflection signals recorded at The Ohio State University Radio Observatory may have been caused by a satellite when it was in or near the southern auroral zone.

Observations have been reported which indicate a strong correlation between the occurrence of 20-mc signal bursts of WWV as observed at Columbus, Ohio, with passes of Sputnik III through the northern auroral regions.[4] The data are summarized in Fig. 1 which shows the WWV enhancement, to an arbitrary scale, as a function of time before and after near approach of the satellite to Columbus for 3 to 4 consecutive passes per night (between local midnight and 6 A.M.), on 14 nights during January and February, 1959. A total of 50 passes is involved. The enhancement of WWV shows a peak at the time of near approach of Sputnik III, but larger peaks somewhat before and after. It is considered significant that these peaks occurred when, on the average, the satellite was passing through the northern horns of the outer Van Allen belt and the northern auroral zone. Other evidence suggests that these enhancements may be due to encounters of the satellite with regions of high-particle density and/or energy in these localities and, further, that these encounters result in the scattering of particles into ionized clouds or streams which travel closer to Columbus with reflections from these clouds producing many of the observed signal enhancements.[4] A theory to account for the ionization build-up around a satellite, and also the detachment of the ionized region to form a traveling cloud or satellite ghost, has been advanced by Singer and Walker,[5,6]

Referring again to Fig. 1, the three largest peaks occur close to the time of near approach. The next largest peak occurs 50 minutes before near approach, when the satellite was, on the average, near the south auroral zone. Although the peak is not as large as the others, it hints at a southern hemisphere effect.

Further evidence of this effect is provided by Fig. 2 which shows the tracks of Sputnik III between 8:18 P.M. EST on August 9, and 7:51 A.M. EST on August 10, 1959, during which time the satellite completed seven successive trips around the earth. All WWV enhancements observed at The Ohio State University Radio Observatory on 20 mc are shown on the tracks as blocks or wedges. Actually, the receiver response band was set for approximately 20.001 to 20.003 mc, so that all WWV enhancements recorded involved some Doppler shift in frequency. The blocks or wedges indicate both duration and relative strength of the enhancements. All enhancements recorded during the eleven and one-half hour interval are shown, none being omitted. In addition, a symmetrical burst sequence reported by Baron and Lerner is shown by the shaded wedges.[7] This sequence was recorded by the National Bureau of Standards at Boulder, Colorado on 15 mc. The signals of WWV (Washington, D. C.) were those recorded, and it is interesting to note that this sequence is very symmetrical with respect to a point midway between Washington and Boulder. Other examples of symmetrical burst sequences have been reported earlier.[8]

A sudden and complete drop out of the Sputnik III transmission on 20.0045 mc lasting 40 seconds was also noted on the fourth pass at the time the satellite was crossing the northern auroral zone. This is indicated in Fig. 2 by the square box over Hudson Bay. This was the pass on which the strongest WWV enhancements were noted at Columbus. Such drop outs, in association with a WWV enhancement, may be indicative of the appearance of a cloud of sufficiently high electron density between the satellite and Columbus to cut off the satellite signals while simultaneously providing a large echo area for the reflection of WWV.

Of particular interest in Fig. 2 is the concentration of the WWV enhancements

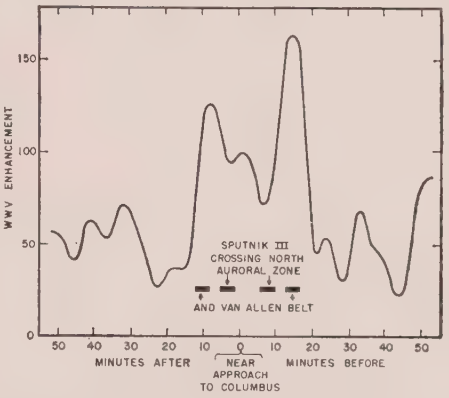

Fig. 1—Enhancements of WWV 20-mc signals recorded at the Ohio State University Radio Observatory as a function of the time before and after the near approach of Sputnik III for 50 passes on 14 nights during January and February, 1959.

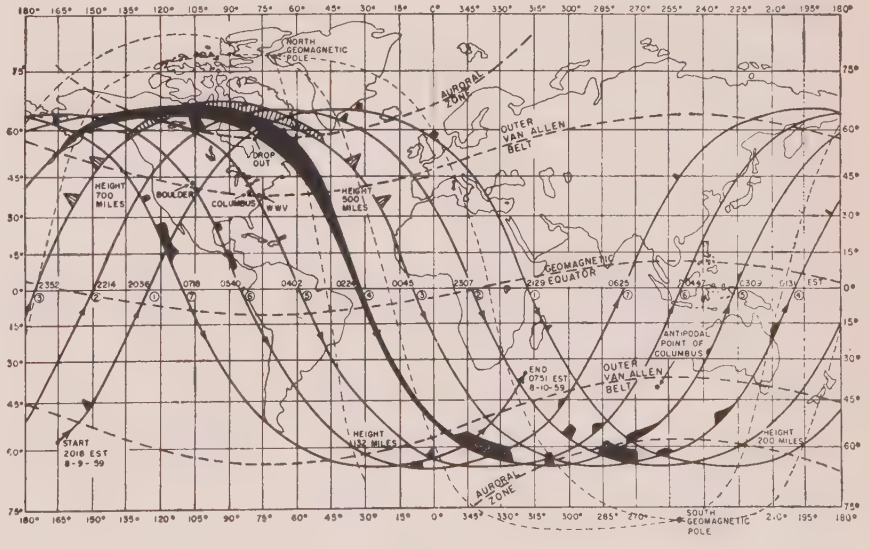

Fig. 2—Enhancements of WWV 20-mc signals recorded at the Ohio State University Radio Observatory, Columbus, as a function of the position of Sputnik III on the night of August 9–10, 1959. The enhancements appear as the solid blocks or wedges on the satellite tracks. Shown also is a symmetrical burst sequence of the WWV 15-mc signals recorded at Boulder, Colorado, as reported by Baron and Lerner (shaded blocks or wedges).

* Received by the IRE, April 18, 1960.

[1] J. D. Kraus, "Detection of Sputnik I and II by CW reflection," PROC. IRE, vol. 46, pp. 611–612; March, 1958.

[2] J. D. Kraus, R. C. Higgy, and J. S. Albus, "Observations of the U. S. satellites Explorers I and III by CW reflection," PROC. IRE, vol. 46, p. 1534; August, 1958.

[3] J. D. Kraus and E. E. Dreese, "Sputnik I's last days in orbit," PROC. IRE, vol. 46, pp. 1580–1587; September, 1958.

[4] J. D. Kraus, R. C. Higgy, D. J. Scheer, and W. R. Crone, "Observations of ionization induced by artificial earth satellites," *Nature*, vol. 185, pp. 520–521; February 20, 1960.

[5] S. F. Singer and E. H. Walker, "Wake of a charged body in an ionized gas," *Bull. Am. Phys. Soc.*, vol. 5, p. 47; January, 1960.

[6] E. H. Walker and S. F. Singer, "Wake of a charged body moving in a plasma," *Bull. Am. Phys. Soc.*, vol. 5, p. 234; April, 1960.

[7] R. Baron and L. Lerner, University of Chicago, Chicago, Ill., private communication, February 19, 1960.

[8] J. D. Kraus, R. C. Higgy, and W. R. Crone, "The satellite ionization phenonenon," PROC. IRE, vol. 48, pp. 672–678; April, 1960.

near the northern and southern auroral zones and the outer Van Allen belts. The long tail crossing the equator on pass 4 after the strong enhancement near the north auroral zone may be due to residual ionization which persisted in the north after the pass. The remaining enhancements, and especially the large ones, are concentrated near the north and south auroral regions, with only a few enhancements, which are also weak and of relatively short duration, in the equatorial regions. The probability that such a distribution would result from a random phenomenon is small. Conversely, the probability is large that the distribution is related to Sputnik III.

If this evidence is taken to indicate an interhemisphere effect, two possible explanations come to mind. One is that the effect is caused by antipodal reflection of the radio signals from WWV. The other is that charged clouds or ghosts induced by the satellite according to the theory of Singer and Walker are trapped by the earth's magnetic field and travel from the southern to the northern hemisphere along the field lines to the conjugate point in the northern hemisphere after the manner of Störmer[9] and Van Allen.[10] If the point is near enough to Columbus, a WWV reflection may be detected, but if not, the particles may make more trips between hemispheres until they drift close enough to Columbus to give a detectable WWV reflection. The light dashes in Fig. 2 show the latitude and longitude variation of a few magnetic force lines between the geomagnetic poles. Thus, the magnetic conjugate point of 15° west longitude and 60° south latitude (geographic) is east of Nova Scotia and south of Greenland.

In connection with such a mechanism, it should be noted that the position of an enhancement in Fig. 2 indicates where the satellite was when the enhancement was observed at Columbus. Thus, if there is a significant delay between the appearance of satellite-induced ionization in the southern hemisphere and the arrival of the resulting charged cloud in the north, the satellite ionization would have occurred when the satellite was to the left of the enhancements shown by the blocks and wedges in the southern hemisphere. It may be significant that if all enhancements in the southern hemisphere are arbitrarily shifted backward (to the left) 2 or 3 minutes to allow for such a delay, the enhancements will be at least as symmetrical with respect to the southern Van Allen belt and auroral zone as they appear in Fig. 2.

It should be noted that the position of the auroral zones and the outer Van Allen belt shown in Fig. 2 are nominal ones, the latitude of the belt being subject to considerable fluctuation. No information was available as to their actual positions, on August 9–10, 1959.

The author wishes to acknowledge helpful discussions concerning the effects described with Dr. H. C. Ko and Dr. C. E. Nielsen, and assistance in obtaining certain of the data by Prof. R. C. Higgy and by D. J. Scheer and A. G. Herriman. The orbit data used was provided by the Smithsonian Astrophysical Observatory. The work was supported in part by the U. S. Army Rocket and Guided Missile Agency through a contract with The Ohio State University Research Foundation, and in part by the University's Fund for Basic Research.

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[9] C. Störmer, "The Polar Aurora," Clarendon Press, Oxford, Eng.; 1955.
[10] J. A. Van Allen, "Radiation around the earth," *Sci. American*, vol. 200, pp. 39–47; March, 1959.

A Technique for Short-Term Oscillator Stability Measurements*

Introduction

It is common for stable oscillator specifications either to ignore the short-term stability or to give some maximum deviation. Often a more meaningful specification would be useful. When very precise frequency measurements are being contemplated, a meaningful short-term stability specification would indicate attainable precision. Short-term stability is defined as the effect of the short-term random variation of the frequency about its nominal frequency. These short-term variations are not related to the long-term drift.

An rms frequency deviation specification would give a better feeling for the operation of an oscillator than a maximum frequency deviation specification. The disadvantage of the maximum frequency specification is that it is difficult to determine how much of the variations being measured between oscillators can be attributed to each individual oscillator. This means that the unknown oscillator must be measured against a standard oscillator which is so much better than the unknown oscillator that the random variations of the standard oscillator can be ignored. All the variation in the measured frequency is assumed to be due entirely to the unknown oscillator. When the stability of the unknown oscillator becomes comparable to that of available standard oscillators, the maximum frequency deviation specification breaks down.

The following discussion presents a method of obtaining the rms frequency deviations for Case 1, when a standard oscillator is available, and Case 2, when a standard oscillator is not available. For Case 2, three unknown oscillators are required in order to determine the rms frequency deviation of any one. A Case 3 is presented when two oscillators have comparable stabilities and the assumption can be made that the rms deviations are equal.

* Received by the IRE, April 29, 1960.

Case 1. RMS Frequency Deviation when a Standard Oscillator is Available

When a standard oscillator is available, the instantaneous frequency difference between the unknown oscillator and the standard oscillator can be measured. This measurement will vary according to the random variations in the unknown oscillator. To obtain an rms frequency deviation, a large number of measurements are needed.

We make the following definitions:

f_s—standard oscillator frequency,
f_1—unknown oscillator nominal frequency,
δf_1—unknown oscillator random frequency variation,
f_m—measured frequency difference.

The instantaneous frequency of the unknown oscillator is divided between a nominal frequency which does not vary and a randomly varying frequency which is the term of interest (*i.e.*, $f_1+\delta f_1$). The measured frequency difference is

$$f_m = f_1 + \delta f_1 - f_s.$$

The difference between the nominal frequency and the standard frequency can be obtained by averaging the measured frequency difference, *i.e.*,

$$\bar{f}_m = \frac{\sum f_m}{n},$$

$$\bar{f}_m = \frac{\sum (f_1 + \delta f_1 - f_s)}{n},$$

$$\bar{f}_m = \frac{\sum f_1}{n} + \frac{\sum \delta f_1}{n} - \frac{\sum f_s}{n}.$$

If δf_1 is truly random and a sufficient number of data points are available,

$$\frac{\sum \delta f_1}{n} = 0;$$

and since f_1 and f_s are constant

$$\frac{\sum f_1}{n} = f_1,$$

$$\frac{\sum f_s}{n} = f_s,$$

so

$$\bar{f}_m = f_1 - f_s.$$

Then

$$f_m - \bar{f}_m = \delta f_1$$

and

$$\sigma_{\delta f_1} = \sqrt{\frac{\sum (f_m - \bar{f}_m)^2}{n}}.$$

The nominal frequency of the unknown oscillator must not change because of long-term drift during the period of measurement. This frequency drift would invalidate the $\bar{f}_m$ calculation unless the frequency drift is known and each data point is corrected.

Case 2. RMS Frequency Deviation when a Standard Oscillator is not Available

In general, when two oscillators' frequencies are compared, each will have a nominal frequency and each will have a ran-

dom frequency variation. We make the following definitions:

f_1—#1 oscillator nominal frequency,
δf_1—#1 oscillator random frequency variation,
f_2—#2 oscillator nominal frequency,
δf_2—#2 oscillator random frequency variation,
f_{12}—instantaneous measured frequency difference,

$$f_{12} = (f_1 + \delta f_1) - (f_2 + \delta f_2),$$
$$f_{12} = (f_1 - f_2) + (\delta f_1 - \delta f_2).$$

Again when a large number of data points are available,

$$\bar{f}_{12} = f_1 - f_2.$$

Again the long-term drifts of the oscillators under measurement must be considered. If the oscillators are roughly equivalent, they will drift in the same direction at the same rate so the term $f_1 - f_2$ will almost be constant. When good precision is required it might be necessary to investigate the drift rates of each oscillator and make a correction to each data point.

$$f_{12} - \bar{f}_{12} = \delta f_1 - \delta f_2$$

then

$$\sigma_{(f_{12}-\bar{f}_{12})} = \sigma_{(\delta f_1 - \delta f_2)},$$
$$\sigma_{(\delta f_1 - \delta f_2)} = \sqrt{\frac{\sum (\delta f_1 - \delta f_2)^2}{n}},$$

so

$$\sigma^2_{(\delta f_1 - \delta f_2)} = \frac{\sum (\delta f_1 - \delta f_2)^2}{n}$$
$$\sigma^2_{(\delta f_1 - \delta f_2)} = \frac{\sum \delta f_1^2}{n} - \frac{2\sum \delta f_1 \delta f_2}{n} + \frac{\sum \delta f_2^2}{n}.$$

Assuming the random variations in each oscillator are completely independent, the cross-coupling term

$$\sum \delta f_1 \delta f_2 = 0.$$

Therefore,

$$\sigma^2_{(\delta f_1 - \delta f_2)} = \sigma^2_{\delta f_1} + \sigma^2_{\delta f_2}.$$

This procedure provides one equation with two unknowns. By using three oscillators, the following equations can be obtained and the individual rms frequency deviations calculated:

$$\sigma^2_{(f_{12}-\bar{f}_{12})} = \sigma^2_{\delta f_1} + \sigma^2_{\delta f_2},$$
$$\sigma^2_{(f_{13}-\bar{f}_{13})} = \sigma^2_{\delta f_1} + \sigma^2_{\delta f_3},$$
$$\sigma^2_{(f_{23}-\bar{f}_{23})} = \sigma^2_{\delta f_2} + \sigma^2_{\delta f_3}.$$

Case 3. Identical Oscillators

When two identical oscillators are compared in frequency, their random deviation can usually be assumed to be equal. Then

$$\sigma^2_{(f_{12}-\bar{f}_{12})} = \sigma^2_{\delta f_1} + \sigma^2_{\delta f_2}$$
$$\sigma_{\delta f_1} = \sigma_{\delta f_2},$$

so

$$\sigma^2_{(f_{12}-\bar{f}_{12})} = 2\sigma^2_{\delta f_1}$$
$$\sigma_{\delta f_1} = \frac{1}{\sqrt{2}} \sigma_{(f_{12}-\bar{f}_{12})}.$$

Conclusion

The rms frequency deviation obtained from this procedure is a one-σ value. This means that there is a probability of 2/3 that any instantaneous frequency deviation is equal to or less than the rms frequency deviation; also that there is a probability of 99/100 that this instantaneous frequency is within three times the rms frequency deviation (*i.e.*, 3-σ value).

The rms frequency deviation specification provides a good feeling for the short-term frequency stability of an oscillator. This method of obtaining the rms frequency deviation can provide frequency stability information when a standard oscillator is not available.

Contributions by Marvin Perlman are gratefully acknowledged.

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A Note on Scatter Propagation*

The work described in this article was conducted for Florida Micro-Communications Inc. under FCC permit number C-33-156. The purpose of the experiments was to determine: a) variation in fading rates for a tropospheric forward scatter system as a function of antenna elevation; b) variations in frequency selective fading as a function of frequency separation of two sidebands; and c) the effects of antenna beamwidth on path multipath.

The characteristics of the system used in these measurements were: frequency—400 and 800 mc; distance—156 miles; modulation CW, SSB and AM, antennas—32, 28, 10 and 5 foot parabolic reflectors. The atmospheric refractive index in a over water path was measured to be 370 in March. The terminals for the circuit were Key Marathon, Florida, and Jaruco, Cuba.

To determine the variation of fading rates as a function of antenna elevation, the following experiment was conducted. The transmitter antenna was held fixed at the horizon, the receiving antenna was aimed from 0° to 6° above the horizon by mechanically offsetting the fed horn in the receiving paraboloid. At the same time, signal levels were recorded from another antenna, used usually in diversity with the one under observation. Both antennas have equal characteristics. The signal level charts were then analyzed to determine whether a change in fading rates could be observed between both antennas. Due to the power levels, 250 watts, that were used for this test, the system did not have a large margin of fading. As a result, we were not able to carry our observations beyond a dynamic range of fading of 15 db. The frequencies used were 400 and 800 Mc. As the receiving antenna was elevated above the horizon, it became apparent that fading rates were changing. The fade rates recorded on the elevated antenna were higher than those recorded on the horizontal antenna.

The variation of fade rate is given diagrammatically in Fig. 1.

To determine the effects of antenna beamwidth on received multipath components, the following experiment was conducted. One microsecond pulses were transmitted from Marathon, on a 28-foot antenna at 800 Mc. At the receiving end, photographs were taken simultaneously of the output from three antennas, a 32-foot paraboloid, and two other antennas, 10 and 5 feet in diameter. The beamwidths of these antennas were, 2.5° for the 32 footer, and 9.2 and 18° for the 10 and 5 footers, respectively. In the analysis of the photographic data, the amplitude of the multipath echoes was left out. The graph, shown in Fig. 2, takes into consideration only time delays in the received echo. The variation in time delays for echos arriving through different paths did not agree closely with theory. It was predicted from the literature available on the subject that the multipath echoes would vary in the form of $t = 10 \log G$ (antenna gain) as the antenna's aperture, and gain, varies.

Actually, the experimental data showed that on this path the time delays were closer to the expression $t = 15 \log G$, when $G \geq 30$ db.

This variation could perhaps be explained in terms of the extreme turbulence that was observed in this path, over water and in a tropical area.

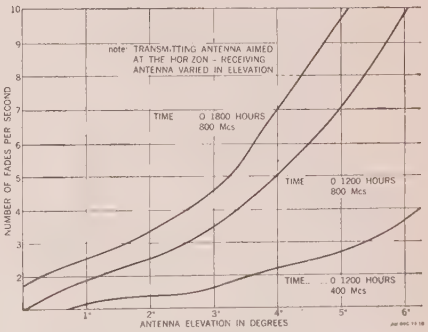

NUMBER OF FADES PER SECOND
ANTENNA ELEVATION IN DEGREES
note: TRANSMITTING ANTENNA AIMED AT THE HORIZON - RECEIVING ANTENNA VARIED IN ELEVATION
TIME 0 1800 HOURS 800 Mcs
TIME 0 1200 HOURS 800 Mcs
TIME 0 1200 HOURS 400 Mcs

Fig. 1—Variation of fading rates with antenna elevation.

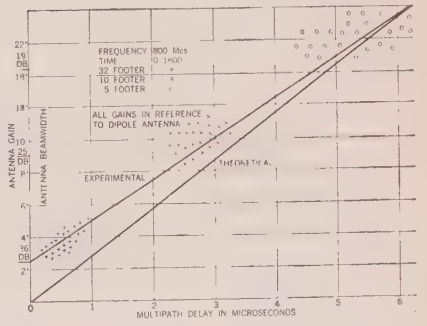
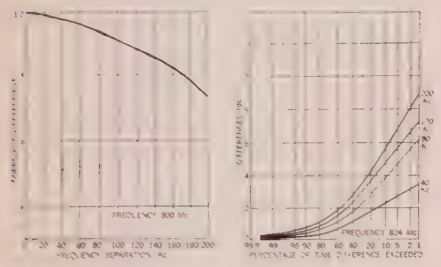


Fig. 2—Multipath delays as a function of beamwidth.

* Received by the IRE, April 6, 1960.

Frequency selective fading was observed in the same circuit by means of recording two signals separated by 40, 80, 120 and 200 kc at a center frequency of 800 Mc. The signals were generated by means of a dual transmitter (actually a converted television transmitter) where the carrier frequency of each transmitter could be offset by a predetermined amount. Each signal was then received independently by two receivers at the Cuban end of the circuit. The noncoherent fading was determined by examination of the recorded outputs of both receivers.

The results of this experiment are given graphically in Fig. 3, both as a function of



(a) (b)

Fig. 3—(a) Correlation of rapid fading vs frequency separation of long range UHF troposphere signals. (b) Selective fading using two frequencies separated 40–80–120–200 kc tropospheric scatter.

noncoherence, and as a function of possible diversity gains.

The author is indebted to Mr. Ventura Montes, Chief Engineer of CMA TV channel 2, Havana, Cuba, and to the Federal Communications Commission for authorizing this test.

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Antenna Pattern Simulation on the Analog Computer*

The purpose of this note is to give a scheme of antenna power spectrum[1] simulation which allows complete generality on the amplitude and phase functions across the antenna aperture. The function to be simulated is

$$P(\theta) = E(\theta)E^*(\theta) \tag{1}$$

where

$$E(\theta) = \int_{-L/2}^{L/2} f(y)e^{j[\beta y \sin\theta + \psi(y)]}dy, \tag{2}$$

and L = antenna aperture, $\beta = 2\pi/\lambda$, $f(y)$ and $\psi(y)$ are the amplitude and phase distributions respectively.

* Received by the IRE, April 21, 1960.
[1] J. D. Kraus, "Antennas," McGraw-Hill Book Company, Inc., New York, N. Y., pp. 348–350; 1950.

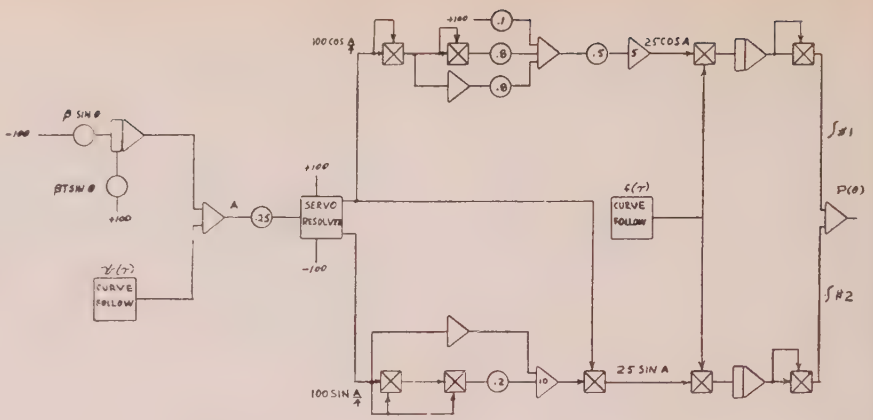

Fig. 1—Antenna pattern simulation. $A = \beta(\tau - T)\sin\theta + \psi(\tau)$.

Performing the multiplication indicated in (1), we have

$$P(\theta) = \left[\int_{-L/2}^{L/2} f(y)\cos \cdot\{\beta y \sin\theta + \psi(y)\}dy\right]^2 + \left[\int_{-L/2}^{L/2} f(y)\sin \cdot\{\beta y \sin\theta + \psi(y)\}dy\right]^2. \tag{3}$$

This function is simulated by holding θ as a parameter, giving it a discrete value for each computer run, out to some chosen θ_0.

Distance across the aperture is made analogous to time: $y \equiv t$. Because the computer does not operate in negative time, the transformation $t = \tau - T$ is made, where $T = L/2$. Eq. (3) then becomes

$$P(\theta) = \left[\int_0^T f(\tau)\cos \cdot\{\beta(\tau - T)\sin\theta + \psi(\tau)\}d\tau\right]^2 + \left[\int_{\cdot\tau}^T f(\tau)\sin \cdot\{\beta(\tau - T)\sin\theta + \psi(\tau)\}d\tau\right]^2. \tag{4}$$

The $\beta(\tau - T)\sin\theta$ term is generated as a ramp function from an integrator, β, T, θ being constants. Time τ is the computer's independent variable. The amplitude and phase distributions, $f(\tau)$ and $\psi(\tau)$, are functions of choice. In this simulation, they are generated by a pair of curve plotters operated in the follower mode. The functions are simply drawn in conducting ink on fairly stiff plastic sheets and the electromagnetic recording heads of the curve followers are made to sweep over them.

The $\beta(\tau - T)\sin\theta$ and the $\psi(\tau)$ are summed, and the sine and cosine of their sum is taken servomechanically. Because the servoresolvers do not have continuous rotation, use is made of trigonometric identities by means of which arguments several times greater than 180° can be continuously resolved. The identities used are $\sin A = \cos(A/4)[4\sin(A/4) - 8\sin^3(A/4)]$ and $\cos A = 8\cos^4(A/4) - 8\cos^2(A/4) + 1$. Thus if A is, for example, of the order of 2π, the servoresolvers have to deal only with angles of the order of $\pi/2$.

Fig. 1 shows a simplified computer diagram of the simulation. The points labeled $\int\#1$ and $\int\#2$ are the first and second terms, respectively, on the right side of (4).

The work of Blumberg[2] contains calculated antenna patterns for some simple shapes of amplitude and phase distributions; uniform, triangular, cosine, and cosine squared. Although there is nothing difficult about such calculations, they are laborious and get quite out of hand when one wishes to introduce higher order departures or asymmetries. They become virtually impossible when one wishes to use completely arbitrary functions or empirical data for the distributions.

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[2] M. Blumberg, "The Effect on the Antenna Pattern of Symmetrical Phase Shifts in a One-Dimensional Aperture," Sylvania Electronic Defense Lab., Mountain View, Calif.

Linear-Slope Delay Filters for Compression*

There exists a considerable range of engineering applications for filters which have a linear-slope group time delay. In a recent issue,[1] Cook has described an application to pulse compression radar systems. Similar work is described by Krönert.[2] Another application, to panoramic receivers, was discussed in a recent IRE advertisement.[3] Pos-

* Received by the IRE, May 2, 1960.
[1] C. E. Cook, "Pulse compression—key to more efficient radar transmission," Proc. IRE, vol. 48, pp. 310–316; March, 1960.
[2] R. Krönert, "Pulse compression: parts I and II," *Nachrtech. Z.*, vol. 7, pp. 148–152, 305–308; April, July, 1957.
[3] Advertisement, "Rapid-frequency scans," Proc. IRE, vol. 48, p. 4A; February, 1960.

sible applications also occur in the area of antenna array information processing.

Contrary to the implications in Cook's paper, the engineer will find very little in the unclassified literature to aid him in the approximation problem associated with this type of filter, although the realization problem is well covered. This letter calls attention to our (unclassified) report on the approximation problem.[4] It should be noted that this report is preceded by five years or more by a large amount of successful work at other laboratories, notably at Bell Telephone Laboratories and Westinghouse, on the design of linear-slope group-delay lumped-parameter networks. Unfortunately, all of this work has been classified and therefore cannot be credited here.

The largest attention in recent years has centered on networks which achieve their linear slope over a moderately narrow percentage frequency range (as opposed to a region starting at zero frequency) and the designs we will discuss are of this "band-pass" type. It turns out that the conjugate critical-frequency cluster contributes rather negligibly to the nonlinearity of delay-slope, even for percentage bandwidths approaching 100 per cent, and thus a "narrow-band" approximation is justified in most instances. Networks with negative slopes (over the controlled delay region) are somewhat more tolerant in this regard.

A series of designs, derived by O'Meara,[4] are summarized in Table I. All-pass designs are assumed in order that a desired amplitude function be independently realizable, although minimum-phase realizations are also feasible, based on the same design data. The x_n values are the real (ω) frequency displacements of the critical frequency clusters, to an arbitrary zero reference, while the σ_n are the imaginary (σ) displacements to the same relative scale. Thus one can choose any desired design center frequency, within the limits of a narrow-band approximation, and the desired bandwidth sets the required frequency scale and hence the (linear) differential delay per staggered aggregate (n tuplet) of all-pass networks. The total required differential delay is usually roughly accommodated by adding a sufficient number of aggregates to realize it; this number may typically range from three or four to several dozen, depending upon the required time-bandwidth product and the network efficiency.

The network efficiency corresponds to the *controlled* (linear) time-bandwidth product per all-pass section and, as may be seen from the table, it increases with increasing number of critical frequency staggerings. This efficiency is roughly analogous to the ratio of delay-time to rise-time per network element in a more usual constant-delay network. The theoretical delay characteristc for the sextuplet (six zero-pole pairs in the upper-half complex frequency plane) is illustrated in Fig. 1. As in all the designs

[4] T. R. O'Meara, "The Synthesis of 'Band-Pass,' All-Pass Time-Delay Networks with Graphical Approximation Techniques," Hughes Aircraft Research Labs., Culver City, Calif., Res. Rept. No. 114; June, 1959. Available upon written request.

TABLE I
A Compilation of Constant-Slope Time-Delay Designs

N = Number of Staggered T's	Efficiency Per Cent	Maximum Relative Delay Error in Per Cent of Linear Range	x_1	x_2	x_3	x_4	x_5	x_6	Linear Range in x (Radians /sec)	Linear Range in $\hat{T}_d$ (Seconds)
			σ_1^{-1}	σ_2^{-1}	σ_3^{-1}	σ_4^{-1}	σ_5^{-1}	σ_6^{-1}		
1	11	±1.55	0						0.25 to 1.02	1.94 to 3.74
			2							
2	21.7	±1.22	0	1.5					0.2 to 3.1	0.8 to 2.68
			1.0	0.62						
3	26.7	±1.05	0	1.0	2.6				0.260 to 4.34	0.880 to 3.34
			1.0	0.65	0.54					
4	39.8	±1.06	0	0.97	2.20	3.70			0.1 to 5.4	0.90 to 4.68
			1.48	1.00	0.755	0.570				
6	57.3	±1.18	0	0.7	1.55	2.57	3.8	5.4	0.1 to 7.35	0.80 to 6.74
			2.1	1.5	1.28	1.10	0.9	0.7		

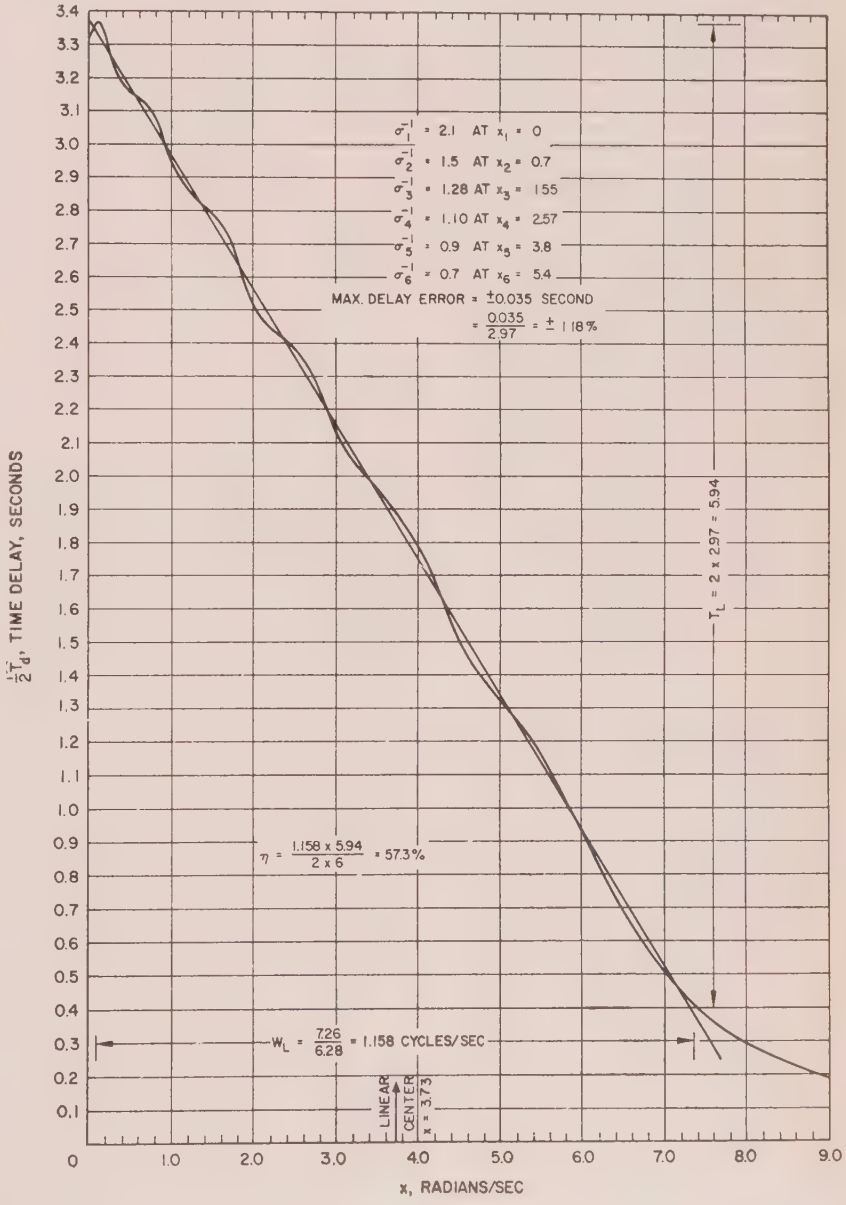
σ_1^{-1} = 2.1 AT x_1 = 0
σ_2^{-1} = 1.5 AT x_2 = 0.7
σ_3^{-1} = 1.28 AT x_3 = 1.55
σ_4^{-1} = 1.10 AT x_4 = 2.57
σ_5^{-1} = 0.9 AT x_5 = 3.8
σ_6^{-1} = 0.7 AT x_6 = 5.4
MAX. DELAY ERROR = ±0.035 SECOND
= 0.035/2.97 = ± 1.18%
$\frac{1}{2}\hat{T}_d$, TIME DELAY, SECONDS
T_L = 2 x 2.97 = 5.94
η = 1.158 x 5.94 / 2 x 6 = 57.3%
W_L = 7.26/6.28 = 1.158 CYCLES/SEC
LINEAR CENTER x = 3.73
x, RADIANS/SEC

Fig. 1—Staggered-sextuplet approximation (N =6).

listed in the table, an approximately equal ripple error characteristic is obtained. The slopes were chosen in an effort to achieve maximum efficiency. A design precision of approximately ± 1 per cent was sought because it is difficult to obtain laboratory adjustments to a much greater precision, with bandwidths of the order of 10 to 30 mc.

An exact technique (in the narrow-band approximation sense) for achieving maximally-flat approximations to any desired group delay function has also been evolved and will be described in a forthcoming report.

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Pulse Shaping with Variable-Capacitance Diodes*

Variable-capacitance diodes offer some advantages over conventional capacitors when used as pulse-shaping elements. The speed at which the junction capacitance responds to a change in diode voltage is extremely fast, on the order of 10^{-10} seconds; so the ultimate frequency limitation of these devices is beyond that required by most present-day applications. The only restriction is that the input signal be kept small so that the diode will never be driven into conduction. In some cases, this may require that a small negative bias be applied to the diode in series with the input; but if the positive peak input pulse is less than +0.5 volt, such a step is not necessary.

The following small-signal analysis derives an analytical expression for a variable-capacitance integrating and differentiating circuit.

The capacitance of a variable-capacitance diode is given by

$$c = \frac{k}{V^n} \tag{1}$$

where k is a constant proportional to the junction area and n is a constant related to the junction impurity distribution. For abrupt junction diodes $n = \frac{1}{2}$, and for linearly graded junction diodes $n = \frac{1}{3}$.

The voltage across a variable capacitance diode as a function of stored charge may be found from (1) by integrating

$$c = \frac{dq}{dV} = \frac{k}{V^n}$$

$$dq = k\,\frac{dV}{V^n}$$

and solving

$$q = q_0 + \frac{k}{1-n}\,V^{1-n} \tag{1a}$$

$$V = V_0 + \left[\frac{(1-n)}{k}\,q\right]^{1/(1-n)}. \tag{2}$$

* Received by the IRE, March 24, 1960.

For a series RC network using a variable-capacitance diode, the transient solution for any input wave function, $f_{(e)}$, may be found by solving

$$f_{(e)} = \left[\frac{(1-n)}{k}\,q\right]^{1/(1-n)} + \frac{dq}{dt}\,r. \tag{3}$$

Case I—Diode Integrator

Consider the circuit shown in Fig. 1.

Assume that the input pulse $f_{(e)}$ has a period t such that $V_r \gg V_c$. Eq. (3) may then be expressed as

$$f_{(e)} = \frac{dq}{dt}\,r$$

and solved as

$$q = \frac{1}{r}\int f_{(e)}dt.$$

Substituting into (2), we get

$$V_c = V_0 + \frac{(1-n)}{rk}\left[\int f_{(e)}\right]^{1/(1-n)}.$$

Thus, we see that a diode integrator operates on an input voltage pulse to provide an output proportional to:

$$\left[\int f_{(e)}\right]^{1/(1-n)}.$$

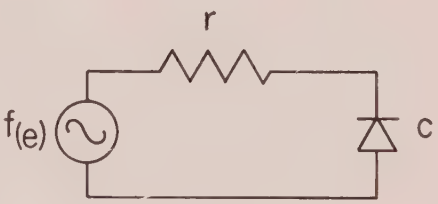

Fig. 1.

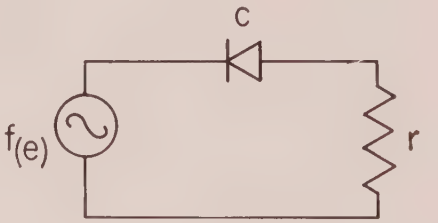

Fig. 2.

Case II—Diode Differentiator

Consider the circuit shown in Fig. 2.

Assume that the input pulse $f_{(e)}$ has a period t such that $V_c \gg V_r$. Eq. (3) may then be expressed as

$$f_{(e)} = \left[\frac{(1-n)q}{k}\right]^{1/(1-n)}.$$

Solving this, we get

$$q = \frac{k}{1-n}\,[f_{(e)}]^{1-n}$$

$$\frac{dq}{dt} = \frac{k}{[f_{(e)}]^n}\,\frac{d}{dt}f_{(e)}.$$

Thus, a diode differentiator provides an output proportional to

$$\frac{1}{[f_{(e)}]^n}\,\frac{d}{dt}f_{(e)}.$$

Figs. 3 and 4 show some typical transient-response voltages for various input waveshapes to a diode integrator and a diode differentiator. Shown also, for comparison, is the output voltage of a conventional integrating and differentiating circuit. The differences are immediately obvious.

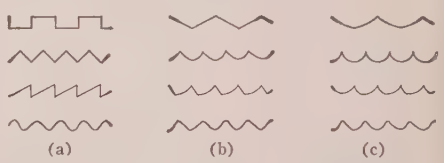

Fig. 3—(a) Input. (b) Conventional integrator. (c) Diode integrator.

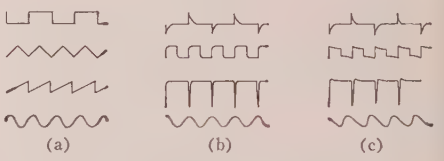

Fig. 4—(a) Input. (b) Conventional integrator. (c) Diode integrator.

It is believed that in many instances the diode characteristic is more desirable than a conventional capacitor.

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Monitoring Paraboloidal Reflector Antennas*

In recent years, very large steerable paraboloidal reflectors have been constructed or planned for radio astronomy and other radio applications, and there is a need for continuous or intermittent monitoring of the shapes of their surfaces. Even after a reflector has been constructed to a true paraboloid, it deflects as it is pointed in different directions, and it is subject to distortion by wind and temperature. These distortions are difficult to investigate; millimeter-wave radar, theodolite survey, stereophotography, and other optical schemes have been suggested.[1] A further need arises in connection with continuous automatic shape correction, as proposed for the 600-foot reflector under construction by the U. S. Navy.[2] As far as is known, no satisfactory answer to these needs has been demonstrated.

In the adjustment of the large cross antenna built at the Radioscience Laboratory, Stanford University, Stanford, Calif.,[3] it is

* Received by the IRE, May 2, 1960. The research reported here was supported by the AF Office of Scientific Res. under contract AF 18(603)-53.

[1] R. M. Emberson and N. L. Ashton, "The telescope program for the National Radio Astronomy Observatory at Green Bank, West Virginia," Proc. IRE, vol. 46, pp. 23–35; January, 1958.

[2] E. F. McClain, "The 600-foot radio telescope," *Sci. American*, vol. 202, pp. 45–51; January, 1960.

[3] G. Swarup and K. S. Yang, "Phase adjustment of large antennas," submitted for publication in IRE Trans. on Antennas and Propagation.

necessary to make accurate phase measurements at 9.1 cm over an aperture of 1339 wavelengths. This is done by placing a gas discharge tube, modulated at 1000 cps, at the feed horn of each of the 32 ten-foot paraboloids in turn. Here it acts as a scatterer for a signal fed into the center of the branched waveguide feeder system, and returns a faint reflection. Because of feeder attenuation and the 50 per cent loss at each branch point both going and coming, the reflection is much less than the small standing reflections caused by acceptable mismatches, but it is identifiable by its modulation, and so its phase can be measured with respect to the signal source.

With a view to application to the adjustment of large paraboloids, an attempt was made to see how far away the discharge could be taken through the open air and still be detectable. A 10-mw *S*-band signal generator was connected to a 3×4 inch horn pointing in the direction of a midget 4-watt standard fluorescent tube placed at a distance of 100 feet. The modulated echo was mixed with a strong local CW signal whose phase was adjusted to obtain a null in a coherent detection system. With an echo strength of −130 dbm, it was found possible to measure displacements of the tube with an accuracy of ±0.1 inch. We do not doubt that better precision and sensitivity would be attainable with attention to details.

However, the precision and accuracy already demonstrated would suffice for a shape-monitoring system along the following lines for even the largest paraboloids now in use. Many discharge tubes would be installed in the parabolic surface. Then a CW signal would be fed into the transmission line near the receiver and the phase distance to each tube in turn would be monitored.

A feature of this scheme is that it uses the same feed system and frequency as are used for normal operation.

Distortions associated with pointing, thermal stress, and wind, which are important at the higher frequencies, but can at present scarcely be investigated, could conveniently be monitored during actual operation. The null technique used at Stanford makes it immediately adaptable to the automatic shape compensation that is under consideration for future giant paraboloids. The mechanical and electrical details are simple and rugged and are suited for all-weather use.

In a further elaboration, surface distortions could be determined in three dimensions by triangulation from additional feed points. A particularly convenient arrangement would be a quarter-wave dipole situated at the vertex of the paraboloid.

Various difficulties arise in trying to measure small phase increments in very long phase paths. The practical arrangements adopted here to cancel out frequency drift and to achieve a null presentation are described in detail elsewhere.[3]

The authors are indebted to Prof. R. N. Bracewell for many valuable discussions.

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On Determining the Polarization Orientation Angle of a Linearly Polarized Source by Analog Techniques*

A method for determining the orientation of the major axis of the polarization ellipse of an incoming signal has been developed that is based on analog computation techniques. Fig. 1 shows a block diagram of this system. Linearly polarized components of an incoming signal are received by a circularly symmetric horn and coupled into the polarization resolver shown in Fig. 2. The three coupling probes spaced 120° apart around the polarization resolver are connected to three identical receivers. Rectified receiver output voltages are connected to the three stator windings of a torque generator. The torque generator is a synchro control receiver and is the analog computing element for the system.

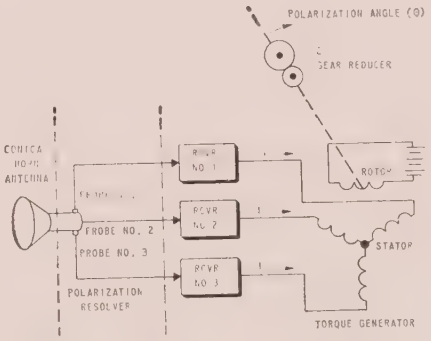

Fig. 1—Analog system for synthesis of polarization orientation angle.

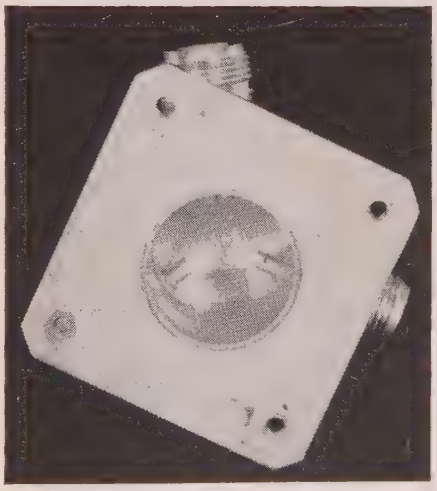

Fig. 2—Polarization resolver.

An analysis of this system has been made for an incoming, linearly polarized signal. The magnitudes of the relative outputs of the three probes are functions of the polarization orientation angle of the incoming signal, being rectified sine waves. The equivalent circuit of the torque generator may be solved for the three currents in the stator windings. Computed relative values of these currents are shown in Fig. 3. The discontinuities in slope shown in Fig. 3 arise from corresponding slope discontinuities in the voltages out of the three probes.

The magnetic fields corresponding to the individual stator currents may be represented as vectors with the proper space orientation, *i.e.*, with 120-degree angular separation. The sum of these three vectors represents the magnitude and direction of the resulting magnetic field in the stator. The rotor is magnetized by a direct current so that the position of the rotor is determined by the orientation of the resultant magnetic field of the stator.

It should be noted in Fig. 3 that the currents repeat for 180-degree change in the polarization angle; consequently, the angular position of the rotor of the torque generator goes through 360° for 180-degree change in the polarization angle. Therefore, a 2:1 gear reduction at the rotor shaft provides an angular position θ corresponding to the polarization angle χ. The angular position θ computed from the currents in Fig. 3 is shown as a function of the polarization angle χ, in Fig. 4. Verification of these results was obtained experimentally by applying the proper voltages to the stator of the torque generator and noting the angle θ. The experimental points are also plotted in Fig. 4 and show very close agreement with the computed values. The maximum error between χ and θ for both computed and measured values of 5° and occurs at odd multiples of 15° of polarization orientation angle. There is negligible error at points which are multiples of 30°.

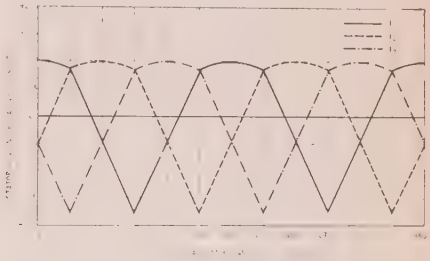

Fig. 3—Calculated currents in torque generator as a function of polarization angle for a linearly-polarized wave incident on the resolver.

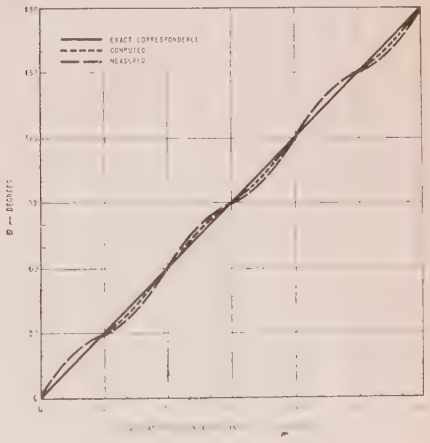

Fig. 4—Angular position θ as a function of polarization angle.

* Received by the IRE, April 21, 1960. The work described in this letter has been supported by the Signal Corps under Contract DA 36-039 SC-78281.

This analysis may be extended to include elliptical polarization of any axial ratio by using the values of V_1, V_2, and V_3 associated with the particular axial ratio.

It is evident that the gains of the three receive channels must be equal, or error will be introduced. The receiver gains must also be controlled by a common AGC system to maintain relatively constant average voltages at the torque generator as the incoming RF signals vary over a large dynamic range. In the case of pulse signals, a peak detector and pulse stretching circuits are necessary to supply the dc voltages to the torque generator.

Another form of this system that eliminates the torque generator has been developed. The three dc voltages may be converted to proportional ac voltages of any convenient frequency, and the computation performed with these ac voltages applied to the stator of a synchro control transformer. An ac closed-loop servo system connected to null out the synchro rotor voltage will generate polarization orientation angle information similar to that generated by the dc torque system. The ac system is more practical because of the advantages of a closed-loop power servo. A breadboard 60-cycle servo system has been constructed which computes the polarization orientation angle for incoming linear polarization.

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Lumped-Model Analysis of Space-Charge Widening*

Lumped models of transistors and diodes have been shown to contribute considerable insight and analytical simplification in the solution of a wide range of problems occurring in the applications of these devices.[1,2] This note is intended to show how the lumped-model approach can be extended to analyze the effects of space-charge layer widening.

Fig. 1 shows a simple low-frequency model of a p-n-p transistor, in which the base region has been represented by the diffusion conductance G_1 and the recombination conductances G_2 and G_3. The quantities p_{ee} and p_{ec} represent excess minority carrier densities at the edges of the two junctions. Then,

$$i_e = p_{ee}(G_1 + G_2) - p_{ec}G_1, \quad (1)$$

$$i_c = -p_{ee}G_1 + p_{ec}(G_1 + G_3). \quad (2a)$$

Solving (1) for p_{ee} and substituting the result into (2a) gives

$$i_c = \frac{-G_1}{G_1 + G_2} i_e + \left(G_3 + \frac{G_1G_2}{G_1 + G_2}\right) p_{ec}. \quad (2b)$$

It is seen that $\alpha_0 = G_1/(G_1+G_2)$; when diffusion predominates over recombination, $G_1 \gg G_2$, and a reasonably high α_0 is obtained. With back-biased collector, $p_{ec} \simeq -p_n$, and with this approximation, the last term of (2b) may be identified as I_{c0}. The dc emitter current[3] I_e is given by

$$I_e = P_{ee}(G_1 + G_2) + p_nG_1 \simeq P_{ee}(G_1 + G_2), \quad (3)$$

the last approximation involving the fact that $P_{ee} \gg p_n$.

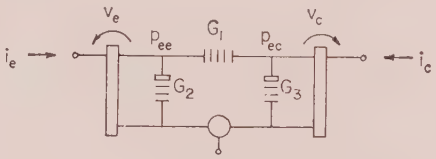

Fig. 1—Lumped model of p-n-p transistor.

The change in the base width w caused by a small change v_c' of collector voltage changes the diffusion and recombination conductances and causes corresponding changes i_c', p_{ee}', and v_e'. Taking v_e and i_c as the dependent variables, the external effects of these changes may be represented by[4]

$$\mu_{ec} = \left(\frac{v_e'}{v_c'}\right)_{i_e=I_e}; \quad g_c = \left(\frac{i_c'}{v_c'}\right)_{i_e=I_e}. \quad (4)$$

Two equivalent approaches may now be used: the parameters of (4) may be calculated by differentiating (1) and (2b) with respect to w, following the general procedure of Early;[5] alternately, the device model of Fig. 1 may be regarded as a varying-parameter circuit in which the G's are functions of v_c. In both approaches, the essential step consists of the realization that the recombination conductances G_2 and G_3 are directly proportional to w, while the diffusion conductance G_1 is inversely proportional to it.[6] This conclusion does not depend on the relative fractions of the base width which are represented by G_2 and G_3. Thus,

$$\frac{-G_1'}{G_1} = \frac{G_2'}{G_2} = \frac{G_3'}{G_3} = m \quad (5)$$

where $G' = \partial G/\partial v_c$, and

$$m = \frac{1}{w}\frac{\partial w}{\partial v_c}. \quad (6)$$

Differentiating (2b) in accordance with (4), and assuming that $\partial I_{c0}/\partial v_c \ll i_e(\partial\alpha_0/\partial v_c)$, we obtain

$$g_c = \frac{-G_1'G_2 \pm G_1G_2'}{(G_1 + G_2)^2} I_e = 2\alpha_0(1 - \alpha_0)mI_e. \quad (7)$$

With the assumptions $p_{cc} \gg p_{ec} = -p_n$, differentiation of (1) yields

$$(G_1 + G_2)\left(\frac{\partial p_{ee}}{\partial v_c}\right)_{I_e} + P_{ee}(G_1' + G_2') = 0; \quad (8)$$

since $G_1 \gg G_2$, one has

$$\left(\frac{\partial p_{ee}}{\partial v_c}\right)_{I_e} = -P_{ee}\frac{G_1'}{G_1} = \pm mP_{ee}. \quad (9)$$

The junction law relates p_{ee} and v_e

$$p_{ee} = p_n(e^{qv_e/kT} - 1), \quad (10)$$

and yields for small increments v_e'

$$p_{ee}' = \frac{q}{kT}(P_{ee} + p_n)v_e' \simeq \frac{q}{kT}P_{ee}v_e'. \quad (11)$$

Combining (9) and (11) there follows

$$\mu_{ec} = m\frac{kT}{q}. \quad (12)$$

Eqs. (7) and (12) are identical to the results obtained by Early.[3]

Alternately, the network of Fig. 1 may be augmented by the addition of incremental conductances proportional to v_c'. Use of (5) leads to the network of Fig. 2. The condition $i_e = I_e$ is imposed in accordance with (4). In computing v_e' and i_c' (to the first order in v_c') one need only consider the currents through G_1 and G_2 due to p_{ee}', and the currents through the incremental conductances due to P_{ee} and $P_{ec} \simeq -p_n$.

$$-i_c' = p_{ee}'G_1 - P_{ee}G_1mv_c',$$

$$i_e' = 0 = p_{ee}'(G_1 + G_2) + P_{ee}(G_2 - G_1)mv_c'.$$

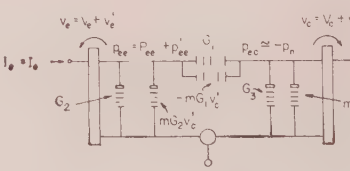

Fig. 2—Lumped model of p-n-p transistor, showing incremental conductances due to space charge layer widening.

Using the approximate expressions in (3) and (11), one obtains

$$i_c' = 2\alpha_0(1 - \alpha_0)mI_ev_c',$$

$$v_c' = \frac{mkT}{q}\left(1 \mp \frac{2G_2}{G_1 + G_2}\right)v_c' \simeq \frac{mkT}{q}v_c',$$

in agreement with (7) and (12).

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* Received by the IRE, April 12, 1960.

[1] J. G. Linvill, "Lumped models of transistors and diodes," Proc. IRE, vol. 46, pp. 1141–1152; June, 1958.

[2] R. N. Beatie, "A lumped model analysis of noise in semiconductor devices," IRE Trans. on Electron Devices, vol. ED-6, pp. 133–140; April, 1959.

[3] We use the notation $i_e = I_e + i_e'$, and so on.

[4] The conductance g_c is evidenced by the nonzero asymptotic slope of the i_c vs v_c characteristic.

[5] J. M. Early, "Effects of space charge layer widening in junction transistors," Proc. IRE, vol. 40, pp. 1401–1406; November, 1952.

[6] As shown in Linvill,[1] the recombination conductance G_r and diffusion conductance G_d of a section of area A and width Δx are given by $G_r = qA\Delta x/\tau$ and $G_d = qAD/\Delta x$.

A Theorem on the Insertion Loss of a Symmetrical Network*

A symmetrical network can be described in terms of its midplane open-circuit impedance Z, and its midplane short-circuit impedance $1/Y$, which are at the same time the impedances of the arms of the equivalent lattice[1] (Bartlett's theorem).

The A matrix of the lattice is

$$\|A\| = \frac{1}{ZY-1}\begin{Vmatrix} ZY+1 & 2Z \\ 2Y & ZY+1 \end{Vmatrix}. \quad (1)$$

Assume input and output terminated by R, then the insertion voltage ratio defined as

$$\frac{2V_{out}}{V_{in}+RI_{in}}$$

can be obtained from the A matrix:

$$V_{in} = \frac{ZY+1}{ZY-1}V_{out} + \frac{2Z}{ZY-1}I_{out}, \quad (2)$$

$$I_{in} = \frac{2Y}{ZY-1}V_{out} + \frac{ZY+1}{ZY-1}I_{out}, \quad (3)$$

$$V_{out} = RI_{out}. \quad (4)$$

By suitable algebraic transformation this can be shown to be equivalent to

$$\frac{2V_{out}}{V_{in}+RI_{in}} = \frac{1}{2}\left\{\frac{Z-R}{Z+R} + \frac{R^2Y-R}{R^2Y+R}\right\}. \quad (5)$$

If R is selected such that $Z=R^2Y$, then from (2) and (3) follows:

$$V_{in} = RI_{in}.$$

The network is matched, and (5) becomes the transfer function of the (matched) symmetrical network:

$$\frac{V_{out}}{V_{in}} = \frac{Z-R}{Z+R}. \quad (6)$$

Even if the network is not matched, both terms on the right side of (5) have the form of transfer functions (6), of matched networks. This permits us to state the following theorem:

> The insertion voltage ratio of a symmetrical 2-port network equivalent to a lattice with arms Z, $1/Y$, and terminated on both ports by R, is equal to the average of the complex transfer functions of 2 matched lattices with arms Z, R^2/Z and R^2Y, $1/Y$, respectively.

The insertion phase and the insertion loss in db are conventionally the phase and 20 times the log of the insertion voltage ratio.

In the special case of the dissipationless network in which Z and $1/Y$ are purely reactive, the transfer functions of the corresponding matched networks become pure phase shifts by $2\phi_1$ and $2\phi_2$, respectively, with

$$\phi_1 = \cot^{-1}\frac{jZ}{R}$$

$$\phi_2 = \tan^{-1}\frac{1}{R\cdot jY}. \quad (7)$$

From simple geometrical considerations it follows that the average of the two complex transfer functions has then the amplitude $\cos(\phi_1-\phi_2)$. Similarly, the phase of the average is $\phi_1+\phi_2$.

The last result corresponds to an expression derived by a different approach by Jones and Cohn[2] for lossless networks only.

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* Received by the IRE, May 9, 1960.

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[2] E. M. T. Jones and S. B. Cohn, "Two theorems for dissipationless symmetrical networks," PROC. IRE, vol. 45, p. 1016, (2); July, 1957.

The Rutile Microwave Resonator*

Single and polycrystalline rutile whose dielectric constant is high[1,2] (about 100) and whose dissipation factor, tan δ[1,2], is low have microwave reflections at the boundary surfaces. A piece of rutile can replace a metal walled cavity because it acts like a microwave resonator.

The fundamental mode[3] inside a rectangular resonator is similar to the TE_{011} mode of a metal cavity with the E and H fields interchanged. The relations between dielectric constant, ϵ', the dimensions, and the resonant frequency were calculated and measurements on the resonant frequency were used to calculate ϵ'. The only possible losses of a perfect crystal resonator will be the coupling and the dielectric losses so that Q_0, the unloaded Q, is given by:

$$Q_0 = \frac{\omega\epsilon'}{\sigma} = \tan^{-1}\delta,$$

where σ is the conductivity.

ϵ' and tan δ were measured[3] in single crystals and polycrystals at different temperatures and at various frequencies from 8 kMc to 50 kMc. It was found that ϵ' increased about 30 per cent as the temperature was lowered from 300°K to 4.2°K. However, at a fixed temperature of 4.2°K, the resonant frequency of the rutile was quite constant.

A typical Q_0 value for the lowest mode at an X-band frequency was 9000 at room temperature. Q_0 also increases by more than a factor of five as the temperature was lowered from 300°K to 4.2°K. At 50 kMc a Q_L, loaded Q, of the order of 10,000 at 4.2°K was easily obtained. Some of the X-band rutile resonators had Q_L's of 85,000[4] at 78°K, but it was found that the Q's vary considerably from one crystal to another. The polycrystalline rutile resonators had Q's of the order of half those of the single crystals.

Paramagnetic spin resonance experiments were simplified considerably by using rutile resonators containing paramagnetic impurities such as Fe^{3+}[5] and Cr^{3+}. Also, the sensitivity[5] of the spectrograph was significantly increased with the use of a rutile resonator.

One of the big advantages of this resonator is the extremely small physical size[3] (for example $0.160\times0.140\times0.40$ in^3 for X band). A resonator about 0.8 cc[5] was used for spectroscopy at 2 kMc. This enabled us to use a 2-inch diameter, 1½-inch air gap magnet to receive the helium cryostat, microwave circuit and mechanical adjustment device for orienting the crystal.

A rutile resonator was also used to observe paramagnetic resonance of powder sample of DPPH which surrounded the crystal in a thin layer. This was possible because the RF field extends outside the crystal for a short distance. A similar procedure could be used for many liquid and powder samples.

In addition to the above virtues, rutile doped with Cr^{3+}[6] and Fe^{3+}[5] has favorable characteristics for use in a maser amplifier. This was confirmed by workers at RCA[7] and M.I.T.[8]

Another application of rutile resonators has been for use as band-pass and band-rejection filters.[9] A VSWR of less than 1.3 was easily obtained within and outside of the pass band in either case. The physical size is much less than that of conventional filters, and the Q and the pass band width can be changed much more readily.

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[1] "Tables of Dielectric Materials," vol. IV, Lab. for Insulation Res., Mass. Inst. Tech., Cambridge, Mass., Tech. Rept. No. 57; January, 1953.

[2] A. Herspring and K. Blank, "Contributions to the phenomenological theory of the dielectric aftereffect in ceramic dielectrics," *Z. Angew. Phys.*, vol. 10, pp. 371–376; August, 1958.

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[4] *Ibid.*, September 16, 1959 to December 15, 1959.

[5] D. Carter and A. Okaya, "Electron paramagnetic resonance of Fe^{3+} in TiO_2 (rutile)," (to be published in *Phys. Rev.*; June 15, 1960).

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[7] C. H. Townes, "Quantum Electronics," Columbia University Press, New York, N. Y., pp. 385–388; 1960.

[8] S. Foner and L. R. Momo, "CW millimeter wave maser using Fe^{3+} in TiO_2," *J. Appl. Phys.*, vol. 31, pp. 742–743; April, 1960.

[9] L. Barash, *et al.*, "Microwave Devices and Measurements, Columbia Radiation Lab. Quart. Rept., Columbia University, N. Y.; December 16, 1959 to March 15, 1960. (Unpublished)

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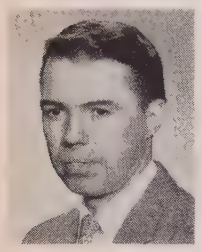

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He served in the U. S. Army from 1943 to 1946, and joined the Radio Corporation of America Electron Tube Division in 1949, where he worked as development engineer on microwave tubes. In 1954 he became engineering leader for Magnetron Design and then manager of Production Engineering in the Color Kinescope Plant, Lancaster, Pa., from 1955 to 1956. He has been with the Power Tube Organization since 1956, and is now manager of the Methods and Process Laboratory, Lancaster, which is responsible for development of new materials and techniques for power tube fabrication and for development of thermionic energy converters.

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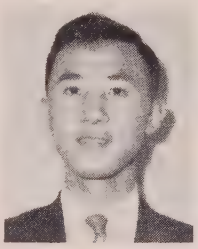

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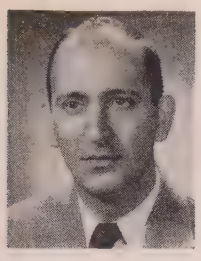

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Karl G. Hernqvist (SM'60) was born on September 19, 1922 in Boras, Sweden. He received the E.E. and the Licentiate of Technology degrees from the Royal Institute of Technology, Stockholm, Sweden, in 1945 and 1951, respectively, and the Ph.D. degree from the same institution in 1959.

K. G. Hernqvist

He was an Assistant Engineer in the Royal Swedish Air Force during 1945 and 1946, where he worked in the fields of radar and microwave instrumentation. From 1946 to 1952 he was concerned with microwave measurements and electron tube research with the Research Institute of National Defense, Stockholm. During 1948 and 1949 he was a trainee under the sponsorship of the American-Scandinavian Foundation at the Radio Corporation of America Laboratories, Princeton, N. J., where he performed research on high-density electron guns. He has been a member of the Technical Staff of the Electronic Research Laboratory at RCA Laboratories since 1952, where he has specialized in the gaseous electronics field. During the past several years he has headed a group working on thermionic converters.

Dr. Hernqvist is a member of Sigma Xi.

❖

Ernest J. Hills was born on January 15, 1903, in Ilvesheim, Germany. He received the M.S. degree in mechanical engineering from the Karlsruhe Institute of Technology, Germany, in 1927.

E. J. Hills

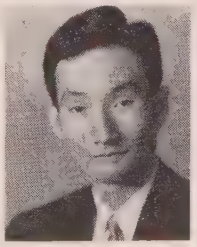
He was then employed in Canada as a design engineer on many projects, such as the design of heating equipment, papermaking machinery, propulsion machinery, machine shop operation, and specialized automatic machine design. In 1948 he continued his work as machine designer and automation specialist in the United States with Union Twist Drill Company, General Instrument Corporation, and Westinghouse, successively. In 1958 he joined the Radio Corporation of America, Lancaster, Pa., as senior engineer in the Power Tube Methods and Process Laboratory.

❖

Nick Holonyak, Jr. (S'51–A'55–M'59), for a photograph and biography, please see page 1506 of the August, 1960, issue of Proceedings.

❖

Hsien Ching Ko (S'53–M'56–SM'60) was born in Formosa, China, on April 28, 1928. He received the B.S. degree in 1951 from the National Taiwan University, China, and the M.S. and the Ph.D. degrees in electrical engineering, both from The Ohio State University, Columbus, in 1953 and 1955, respectively.



H. C. Ko

From 1951 to 1952 he worked on the ionospheric physics and propagation at the Radio Wave Research Laboratories in Taipei, Formosa. Since 1952 he has been associated with the Radio Observatory, Department of Electrical Engineering, The Ohio State University. He has worked on the problems of cosmic radio noise, radio star scintillation, and development of antennas and receivers for radio astronomy. In 1955, he joined the staff of The Ohio State University, where he is presently an associate professor of electrical engineering and assistant director of the Radio Observatory.

Dr. Ko is a member of Sigma Xi, Pi Mu Epsilon, Eta Kappa Nu, the American Astronomical Society, and Commission V of the U. S. National Committee of URSI.

❖

I. Arnold Lesk (S'47–A'52–M'56–SM'59), for a photograph and biography, please see page 1507 of the August, 1960, issue of Proceedings.

❖

Willen D. Meewezen was born in Amsterdam, The Netherlands, in 1925. He obtained the Ir degree in electrical engineering from Delft University, The Netherlands, in 1948.

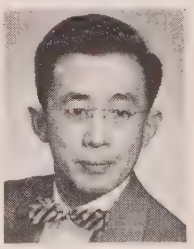
W. D. Meewezen

In 1947, he was appointed a research assistant to the Telecommunications Section of Delft University, where he worked on automatic switching systems until 1948. At that time, he was called into the armed forces and transferred to the Electronic Research Laboratories of the Royal Netherlands Navy; he worked there on the development of new radar equipment. In 1950, he went to Australia, where he joined the Telecommunications Section of Philips Electrical Industries. This is now the Telecommunication Company of Australia Pty. Ltd., and he is currently in the position of Chief Engineer of the Line Telephony Division.

Mr. Meewezen is a member of the Australian IRE and an associate member of the IEE.

❖

Tetsu Morita (S'44–A'49–SM'54) was born in Seattle, Wash., on February 5, 1923. He received the B.S.E.E. degree from the University of Nebraska, Lincoln, in 1944, the M.S. degree in communications engineering in 1945, and the Ph.D. degree in engineering sciences in 1949, both from Harvard University, Cambridge, Mass.



T. Morita

From 1949 to 1953, he was a Research Fellow and Head of the antenna group at Harvard. From 1951 to 1953, he was a consultant on microwave antennas for Trans-Sonics Inc., Bedford, Mass., and in 1953, for the Sylvania Electric Company in Boston, Mass. In 1953, he joined the staff of Stanford Research Institute, Menlo Park, Calif., where he has worked on ECM antennas, missile antenna systems, and an investigation of gaseous discharge phenomena in regard to antennas. He is currently the head of the radiation systems group, Electromagnetics Laboratory.

Dr. Morita is a member of the American Physical Society, Sigma Xi, Pi Mu Epsilon, and the Scientific Research Society of America.

❖

Norris S. Nahman (S'50–A'52–SM'57) was born in San Francisco, Calif., on November 9, 1925. He received the B.S. degree in electronics from California State Polytechnic College, San Luis Obispo, in 1951 and the M.S.E.E. degree from Stanford University, Stanford, Calif., in 1952.

N. S. Nahman

He then entered active service with the U.S. Army where he served as an electronic scientist assigned to the National Security Agency. In 1955, he joined the electrical engineering staff at the University of Kansas, Lawrence, where he is presently the Director of the Electronics Research Laboratory of the Center for Research in Engineering Science. For the past four years he has been primarily engaged in research on nanosecond pulse techniques in the capacity of principal investigator on Project JAYHAWK.

Mr. Nahman is a member of Eta Kappa Nu, Sigma Xi, Sigma Pi Sigma, and Sigma Tau.

❖

Frederick H. Nicoll (A'39–SM'43) was born on June 6, 1908 in Saskatchewan, Canada. He attended the University of Saskatchewan, receiving the M.S. degree in physics in 1931. The following three years were spent in graduate study at the University of Cambridge, England, resulting in the Ph.D. degree in 1934.

F. H. Nicoll

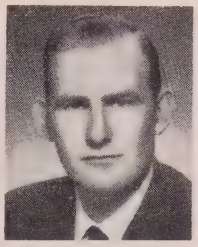
From 1934–1939, he was engaged in television research with a British television laboratory. In 1939 he came to RCA at Camden, where he was engaged in electron optics and cathode-ray-tube design. Since 1942 he has been at the Princeton Laboratories of RCA, being actively involved in the color kinescope program. His recent work has been concerned mainly with display devices. This includes solid-state work on photoconductors and electroluminescence.

Dr. Nicoll is a member of Sigma Xi and the American Physical Society.

❖

John J. O'Connor (S'49–A'52–M'60) was born in Staten Island, N. Y., on December 4, 1926. In 1951 he received the B.S.E.E. degree from Columbia University, New York, N. Y. In 1952 he was a graduate assistant and received the M.S.E.E. degree from the same university.

He was employed by the Ford Instrument Company of the Sperry Rand Corporation, Long Island City, N. Y., as a design engineer, working on analog computers and airborne navigational and radar equipment.



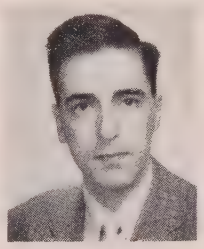
J. J. O'Connor

In 1954, he joined the staff of the Electronics Research Laboratories of Columbia University, where he was concerned with various theoretical and practical aspects of noise in missile systems. In 1959, he joined Hathaway Instruments, Inc., Denver, Colo., as a project engineer, and has been working on various applications of time and frequency standards.

Mr. O'Connor is a member of Tau Beta Pi.

❖

Richard H. Pantell (S'54–A'55) was born on December 25, 1927, in New York, N. Y. He received the B.S. and M.S. degrees from Massachusetts Institute of Technology, Cambridge, in 1950. His studies at M.I.T. were performed within the electrical engineering cooperative course, with four semesters spent at the General Electric Company test program. During the 1950–1951 academic year, he taught electrical engineering at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. From 1951 to 1954, he was a research assistant at the Stanford Electronics Research Laboratory, Stanford, Calif., investigating new techniques for network synthesis. In 1954 he received the Ph.D. degree in electrical engineering from Stanford University.



R. H. Pantell

For the next two years, he was assistant professor of electrical engineering at Stanford University and research associate in the Microwave Laboratory. He taught a graduate course in network synthesis, and did research on the development of a high-power traveling-wave tube. He was granted a leave of absence from Stanford to become a visiting assistant professor at the University of Illinois, Urbana, during 1956–1957. There he worked on the generation of millimeter and submillimeter wavelengths, and taught courses in network synthesis and microwave measurements. He has resumed his position at Stanford and is continuing his research into the generation of microwave energy. He is presently an associate professor of electrical engineering.

Dr. Pantell is a member of Sigma Xi.

❖

William E. Scharfman (S'50–A'55) was born in New York, N. Y., on May 29, 1932. He received the B.E.E. degree from New York University in 1953. He received a Westinghouse Fellowship from The Ohio State University, Columbus, obtaining the M.S.E.E. degree in 1954. The subject of his thesis was in the field of radio astronomy.

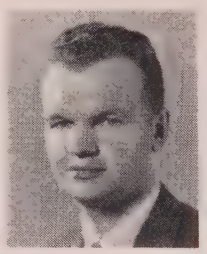
In 1954, he joined the staff of Stanford Research Institute, Menlo Park, Calif., where he has been engaged in the development of ECM antennas, ADF input circuits studies, the development of guidance antennas for missiles, and an investigation of gaseous discharge phenomena in regard to antennas. He is currently a research engineer in the radiation systems group, Electromagnetics Laboratory.

W. E. Scharfman

Mr. Scharfman is a member of the Scientific Research Society of America and Eta Kappa Nu.

❖

W. W. Siekanowicz (A'49–M'55) was born on January 3, 1928, in Lwow, Poland. He received the B.S. degree from the Imperial College of Science and Technology, London University, England, in 1948; the M.S. degree from Columbia University New York, N. Y., in 1950; and the D.E.E. degree from the Polytechnic Institute of Brooklyn, N. Y., in 1960.



W. Siekanowicz

He has been employed in the Tube Division of RCA at Harrison, N. J., since July, 1950, and transferred to the Princeton branch in 1956, where he is now group leader in high-power electronics. He has specialized in the fields of traveling-wave tubes, klystrons and electron beams.

Dr. Siekanowicz is a member of Sigma Xi.

❖

Alan Sussman was born on February 28, 1933, in New York, N. Y. He received the B.S. degree in chemistry from Rensselaer Polytechnic Institute, Troy, N. Y., in 1954, and the Ph.D. degree in physical chemistry from Yale University, New Haven, Conn., in 1957. He did undergraduate work in surface chemistry, specifically oxidation studies of powdered molybdenum disulfide, and graduate work in polymers, radiotracer techniques, and self-diffusion studies in solid ion-exchange polymers.

A. Sussman

Dr. Sussman joined RCA Laboratories, Princeton, N. J., in August 1957, and has worked on image intensifier panels, especially two-color input, two-color output light amplifiers. He is currently studying hot-electron emission from magnesium oxide.

Books

Electronic Engineer's Reference Book, L. E. C. Hughes, Ed.

Published (1960) by the Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. 966 pages+18 index pages. Illus. 5×8. $18.00.

This book is highly recommended as a useful addition to any electronic engineering reference library.

The first edition of this book was exhausted within a year, at which time this second edition was prepared. The publisher and editor are to be commended for their willingness to update their reference work so rapidly. If this policy is continued, this reference can remain as current and timely as the expanding periodical literature, a virtue that so far has been claimed by few.

Good judgment has been shown by the manner in which the book has been organized. The book is comprised of a large number of articles written by approximately 75 experts. Each article bears a symbol to identify the author. One section contains concise biographical sketches of the authors which include their major technical publications and industrial affiliations. A few of the articles are excerpts taken directly from commercial technical literature. References to technical periodicals are integrated into each article near the material to which they relate, which is very desirable. The articles are assembled into 9 subdivisions each comprised of up to 9 articles. The titles of the subdivisions are Fundamentals, Radiations, Electrics, Valves, Materials, Vibrations, Computing, Automatics, and Miscellaneous. The subdivision titles are vague because more specific titles would fail to describe the broad coverage of each subdivision. Introductory to the 9 subdivisions are 23 pages devoted to discussions of the history of electronics, human sensory characteristics that may be numerically defined, the scientific method, and pure vs applied science.

The topics have been chosen from a widely divergent field of disciplines in an attempt to bring together in one reference work material that is not readily available to the electronics engineer. The broad coverage brings into sharp focus the fact that the electronics engineer may no longer confine his activities to a narrow specialty if he is to be truly creative and participate in the advance of the technical frontiers in this space and atomic age.

The editor has prescribed the limits of the book to benefit particularly the engineer whose output is used for industrial purposes. Material peculiarly applicable to the design of military electronic equipment has been intentionally omitted. He has assumed that the subjects of radar, radio, television, communications, filters, and maintenance testing equipment have been exhaustively covered elsewhere, and therefore he has omitted all but the briefest reference to these subjects.

Many of the materials, devices, and instruments are identified by their European trade names with which American engineers are unfamiliar. This and the scarcity of those items in this country limit the value of some of the technical information in the book.

Even though this book is not geared to the specific needs of the telecommunication or military electronic equipment designer, they can profit from the broadening influence that exposure to the book can produce.

On the whole, this reviewer found the book to be a difficult job well done.

GUSTAVE SHAPIRO
National Bureau of Standards
Washington, D. C.

The Cathode Ray Tube and Its Applications, 3rd Rev. Ed., by G. Parr and O. H. Davie

Published (1959) by Reinhold Publishing Corp., 430 Park Ave., N. Y. 22, N. Y. 422 pages+11 index pages+xii pages+bibliography by chapter. Illus. 5¾×8¾. $9.50.

To cover, in one volume, the vast area of modern oscilloscope applications seems almost impossible. One might expect to find either width or depth, but not a combination of both.

Yet, this has been accomplished by G. Parr and O. H. Davie in the new (1959), revised and enlarged, third edition of their well known book: "The Cathode Ray Tube and its Applications." The topics covered in the text are presented with clarity and with attention to detail, thus giving the reader a thorough understanding of basic principles. For further reading and study, each chapter closes with a literature reference, listing, on the average, some 40 items per section, most of them up to date until about 1958.

The work is divided into 16 chapters. The first two deal with the cathode ray tube itself, and the next six with the basic circuits directly associated with the tube. These include power supplies, video amplifiers, sweep circuit techniques, special time bases, calibrators and switching devices, and frequency time bases.

The important subject of photographic recording is treated in a special section.

The seven remaining chapters are devoted to applications of the oscilloscope in many areas of engineering and science. This part of the book is ushered by a chapter on transducers. In the following section on mechanical engineering, we find such items as ultrasonic flaw detection, engine indicators, and seismology. The electrical engineering section includes high voltage testing, FM-AM monitoring, radar indicators and television, including color. The closing two chapters are devoted to particle physics and to electromedical applications.

The book is written in a clear and concise language and is very well printed and illustrated. With its wealth of bibliography, it is an excellent text and reference book. In this reviewer's opinion, this is a "must" for every worker in the field!

KURT SCHLESINGER
General Electric Co.
Electronics Park
Syracuse, N. Y.

Self-Saturating Magnetic Amplifiers, by G. E. Lynn, T. J. Pula, J. F. Ringelman, and F. G. Timmel

Published (1960) by McGraw-Hill Book Co., Inc., 330 West 42 St., N. Y. 36, N. Y. 213 pages+3 index pages+xv pages+bibliography by chapter. Illus. 6¼×9¼. $8.00.

A practical approach to understanding the principles and design techniques of self-saturating magnetic amplifiers is given. The presentation moves smoothly from an introduction based upon transformer fundamentals, known to any graduate engineer, through a brief study of solid-state component property requirements to amplifier circuit behavior. All of the common circuit configurations are treated in detail with due consideration given to the subtleties of operation which are responsible for over-all circuit behavior and transient response. The treatment of amplifiers in closed loop systems is adequate to relate magnetic amplifier behavior to other amplifying devices used in servomechanisms. Practical data are presented to facilitate reactor design.

The book is written for a graduate engineer interested in applying a magnetic amplifier to a particular situation. The text is decidedly slanted more towards proportional control amplifiers than digital control amplifiers, although most of the significant principles are presented. The treatment is somewhat sketchy in some areas due to the size of the work. However, this is not detrimental to the purpose of the book, and a fairly complete bibliography affords the interested reader the opportunity of further study.

The book is the clearest general treatment of self-saturating magnetic amplifier principles and design read by this reviewer. It fulfills a definite need for such a text and should be welcomed by all newcomers to the field.

H. A. PERKINS, JR.
Philco Corp.
Willow Grove, Pa.

An Introduction to Linear Programming and the Theory of Games, by S. Vajda

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 63 pages+2 index pages+ix pages. Illus. 5⅜×8¾. $2.25.

"Linear Programming" and "Theory of Games" carry an aura of mystery in present-day technology. The uninitiated should welcome a brief description of what these subjects have to offer. The thin volume under review appears suited to fulfill this purpose for those who must read on the run. Tersely presented examples illustrate the kinds of problems which can be attacked and the methods which have been developed for their solution.

We learn from the book that linear programming deals with situations in which costs, profits, production rates, and the like, obey linear relationships within some bounded domain. Since there are inescapable realities, such as the finite capacity of machines and the inability to manufacture

negative quantities of a product, the linear equations must be supplemented by a set of constraints in the form of inviolable inequalities. To lay out a program which minimizes cost or maximizes profit within this framework, we must perform some cut-and-try analysis. Linear programming theory teaches us how to systematize the attack, explains various effective procedures which have been discovered, tells us what to do when we reach a dead end in our trials and, in keeping with the spirit of present times, shows how to make computing machines do most of the work.

In game theory, matters are complicated by malevolent competitors intent on making their own profits at the expense of ours. Both they and we have a choice of strategies and, since the enemy strategy is unknown, we cannot be sure of attaining the best possible result in any single event. We can, however, provide against the worst the enemy might do to us. Also, by a properly chosen mixed strategy, we can maximize our expectation of gain against an opponent who also is free to mix his moves.

The selection of worked examples is sufficient to get the reader off the ground, and follow-up references are furnished. Detailed proofs are given of the duality theorem in linear programming, which shows the relation between certain corresponding maximizing and minimizing problems, and of the main theorem on existence of a solution in the theory of games.

W. R. Bennett
Bell Telephone Labs.
Murray Hill, N. J.

Health Physics Instrumentation, by John S. Handloser

Published (1959) by Pergamon Press Inc., 122 E. 55 St., N. Y. 22, N. Y. 177 pages+5 index pages+x pages. Illus. 5½×8¾. $6.50.

Mr. Handloser has given the reader an appreciation of the complexities of measuring the biological effect of nuclear radiation. In instrumentation for health physics, as this measurement science is called, most users are interested in a single quantity—the relative biological effect—whereas the nuclear radiation itself is multidimensional in nature.

First, radiation can be divided into four common categories (alpha or beta particles, gamma rays, and neutrons), all of which have highly different characteristics and methods of interacting with matter. Second, in practice one seldom encounters a single kind of radiation; mixed fields are more likely. Third, each radiation has an energy distribution spectrum, as well as an intensity. And last, the radiation may emanate from a nuclear machine, a radioactive gas, from surface dirt, from airborne dust, or from cosmic space. In the presence of these variables, it is nigh impossible with a single instrument to measure the net biological effect, and certain prior knowledge about the radiation is necessary to the measurement.

The book is divided into nine chapters. The first (Introduction) is devoted to the bases of health physics, to types and characteristics of radiation, definitions, units, and limits of radiation. The second chapter discusses the nature and operation of the various radiation detectors commonly used. Portable survey instruments, fixed monitors and neutron monitors are described in the next three chapters. Chapters 6 and 7 are devoted to counting rooms and equipment, and to calibration techniques—important topics in a successful health physics program. The book closes with two chapters titled miscellaneous instrumentation and allied equipment. In addition to a general reference list, each chapter concludes with a well-selected bibliography.

This reviewer considers the book more of a survey than a textbook. Understandable by anyone with a reasonable aptitude for physics, it is more useful to a reader already equipped with some knowledge of particle physics, and of the specialized electronic techniques in nuclear instrumentation. The author has selected a descriptive rather than expository method of treatment, in which much data, formulas, and other facts are given without derivation or explanation.

In a book of this length (177 pages), it is impossible to do justice to all facets of a discipline as complex as health physics. In addition to addressing himself to the major topic, the author has attempted to cover the broader electronic principles of nuclear instrumentation as needed in health physics. Here the treatment is spread rather thin, as is especially evident in the last chapter, and in the two-page discussion of statistical effects. A serious reader will be forced to other references on nuclear electronic instrumentation. (Unfortunately, there are no comprehensive up-to-date sources.) The space devoted to descriptive discussion of electronic techniques, for example the discussion on neon glow lamps, could have been better devoted to deeper expository treatment of health physics topics.

The book would be improved by some attention to historical trends, to newer principles and devices for health physics monitoring (*e.g.*, glass dosimeters), and to some predictions of the future that lies ahead. In spite of these criticisms, Mr. Handloser has made a valuable contribution by compiling the problems, devices and techniques that constitute health physics.

A. B. Van Rennes
United Research, Inc.
Cambridge, Mass.

Nonlinear Electrical Networks, by William Lewis Hughes

Published (1960) by the Ronald Press Co., 15 E. 26 St., N. Y. 10, N. Y. 296 pages+3 index pages+11 appendix pages+x pages. Illus. 6¼×9¼. $10.00.

Although primarily a text for graduate electrical engineering students, this book should also serve as a useful reference for research and development engineers and physicists. Its title might well have been prefixed by "An Introduction to . . . " or "Fundamentals of . . . " in the sense that its coverage is broad and primarily concerned with classifying types of problems and approaches to their solution, rather than presenting a detailed analysis of a few specific problems. In other words, the author has presented the material in a form that is applicable to almost any nonlinear circuit problem that may arise. Having pointed the way, he either takes the reader through significant details of a well-chosen example, or cites a pertinent reference that will offer equivalent enlightenment.

The book opens with a chapter concerning the definition and nature of basic nonlinear electrical elements, comparing their terminal characteristics with those of more conventional linear devices. Chapter 2 is devoted to the simpler class of nonlinear circuits, namely, circuits without energy-storage elements (inductances or capacitances). In preparation for the material to follow, Chapter 3 presents a review of pertinent mathematical techniques including multiple-variable power series, trigonometric multiple products, Fourier series, algebraic solution of polynomials, and curve fitting. Chapters 4 through 12 finally discuss the more difficult class of nonlinear circuits that contain energy storage devices.

More specifically, Chapter 4 makes use of simple energy storage circuits to illustrate such concepts and methods of analysis as exponential series, complex convolution, reversion, iteration, and equivalent linearization. Chapters 5 and 6 discuss steady-state and transient performance of more complex nonlinear networks, with particular emphasis on iterative and iterative-averaging methods. Chapters 7 and 8 are concerned with network stability, which for nonlinear circuits is far more than a mere extension of the concepts of linear circuit stability. Chapters 9 and 10 conclude the book with a discussion of nonlinear oscillations and the generation of integral and integral-ratio subharmonics.

The book does not pretend to present a single, all-encompassing, "cut-and-dried" method for the solution of nonlinear network problems. On the contrary, it discusses quite a number of methods that vary not only with the problem, but also in rigor. Actually, the author is to be complimented for exploring several areas of the subject that are not amenable to formal proof, but may be attacked by "organized speculation." In these cases, he combines experimental evidence with mathematics to yield results that are approximate, but useful. While some may not approve of this approach, this reviewer feels that such results are infinitely better than none at all, especially when the limitations of the analysis are frankly and clearly stated.

Like nearly every first edition, the text does contain errors. However, errors noted were not mistakes in principle, but mere slips in exponents, subscripts, symbols, equation numbers, etc. Although a casual reading indicated that most chapters were error free, one chapter was noted to contain five minor errors (subscripts in equations 4-71, 4-72, and 4-73; equation reference following equation 4-98; and symbol in equation 4-177). Although not a serious need, a brief list of corrections could well accompany each copy of the first edition.

Considering all its features, this book represents a better than average treatment of the subject. It should be well received among educators as well as practitioners.

H. M. Straube
RCA
New York, N. Y.

Cybernetics and Management, by Stafford Beer

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 208 pages+3 index pages+3 bibliography pages+xviii pages.

"The beginning is half of the whole." Like this dedicatory quotation from Pythagoras in the original language, much of Mr. Beers' fascinating and altogether solid work on cybernetics, "the science of communication and control," will be Greek to most people in responsible line management positions for whom it is ostensibly titled. But to members in several IRE Professional Groups, and especially to dedicated management advisers and consultants, it is of particular moment.

Developed by its English author from a 1957 invited lecture before the Royal Swedish Academy for Engineering Sciences and an immediately subsequent series of seminars and conversations, it has a naturally English and European emphasis as to sources and technique; but the matter is universal in substance, and American sources are not neglected.

Ranging functionally in both breadth and depth over the realms of computer logic, operations research, systems engineering, statistical communication theory and management science, with excursions into the provinces of philosophy and ethics, the whole is still connectively treated, and brilliantly so, in no more than 214 pages.

The very organization of this book is logical as well as novel. Divided into four formal parts, each introduced by a systematic "Connective Summary," the twenty chapters are no less aptly titled than are the parts: "I. Basic Notions; II. The Logical Theory of Cybernetics; III. The Biophysical Theory of Cybernetics; IV. The Analogue Theory of Cybernetics"—concluded by a quite comprehensive listing of 40 references and an excellent combination topical and blue-book name index ranging from Aristotle through "possibly the greatest logician of the middle ages," the 14th Century English Franciscan, William of Ockham, to Wiener, Von Neumann, Shannon and other modern computer and information theory giants. (In charity we may presume the still earlier 13th Century great exponent of Aristotelean logic, Thomas of Aquinas, was bypassed because he happened not to be an Englishman!)

Categorical restriction of cybernetics to binary notation and logic (Chapter X), albeit with passing acknowledgment of Von Neumann's "majority" concept (Chapter XI), seems premature in the present state of the art. Of the sparse illustrations, the two in Chapter IX are interchanged in sequence, probably because they are untitled and unnumbered. But these are negligible criticisms.

This is a stimulating and sometimes provocative book, but hardly one for casual reading. (Its review has been a job of work requiring frequent resort to references, including Webster.) But to one seriously involved in this automation era, including the military as well as the civilian expert, it is commended for possession and use. "Cybernetics and Management" may well serve to aid accomplishment in America, as much as in England, of that necessary objective, survival, as the author emphasizes in his preface: "We desperately need some radical new advance, something qualitatively different from all our other efforts, something which exploits the maturity and experience of our culture. A candidate is the science of control." To which we say a fervent, "Amen!"

James J. Lamb
Ramo-Woodridge
Div. of Thompson Ramo Wooldridge, Inc.
Sierra Vista, Ariz.

Automatic Language Translation, by Anthony G. Oettinger

Published (1960) by Harvard University Press, Cambridge 38, Mass. 351 pages+4 index pages+xix pages+9 bibliography pages+15 appendix pages. Illus. 6¾×9½. $10.00.

This book is a useful addition to the literature on automatic language translation. However, its title and subtitle (Lexical and Technical Aspects, with Particular Reference to Russian) are misleading, since very little attention is given to translation, even less to language, and the fundamental semantic problems of lexicography are barely mentioned and not discussed. This stems from Professor Oettinger's ". . . conviction that an operating automatic dictionary is a necessary basis for effective, linguistically significant research on syntax and semantics, as well as a necessary first approximation to an automatic translator." (Preface, p. vii.) The book is essentially a detailed and comprehensive description of an operating automatic Russian-English dictionary.

The first three chapters (Automatic Information-Processing Machines, The Structure of Signs, and Flow Charts and Automatic Coding) provide an excellent introduction to the basic content of automatic information-processing machines and their use. A very happy balance between detail (relevant to the Univac) and generality is achieved. Chapter 4 (The Problem of Translation) is superficial and nonilluminating. Very little distinction is made between translation from one natural language to another and "translation" from one representation of a given natural language to some other representation of the same language. The next five chapters (Entry Keys for an Automatic Dictionary, Morphological and Functional Classification of Russian Words, Dictionary Compilation, Dictionary Operation, and Problems in Dictionary Compilation and Operation) give an exhaustive account of the Harvard Automatic Russian-English Dictionary. The goal of the maximum possible elimination of the need for human intervention, not only in dictionary operation but in dictionary compilation as well, is very impressively achieved. The practical problem of error detection and elimination is given the recognition it deserves. The approach throughout is strongly machine-oriented and it remains to be seen whether elegance in compilation has been bought at the expense of applicability of the dictionary to automatic translation and to syntactic and semantic investigations. The final chapter (From Dictionary to Translator) consists of very brief and sketchy comments on topics ranging from automatic affix interpretation to automatic discovery of translation algorithms.

The book is well organized and very clearly written. There is nothing obscure or ambiguous in the presentation. A tremendous amount of necessary detail is included, and at the same time, the relation between particular details and the over-all system (the Automatic Dictionary) is kept clearly in view. It is unfortunate that the relation between the automatic dictionary and the over-all process of automatic translation is not similarly kept clearly in view.

This book will be of interest to anyone concerned with the automatic processing of language material, whether for translation, information retrieval, or linguistic research.

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Information Processing

Published (1960) by UNESCO, distributed by International Documents Service, Columbia University Press, 2960 Broadway, N. Y. 27, N. Y. 466 pages+21 index pages+xxxi pages+translations in English, French, Russian, German, and Spanish. Illus. 8¼×12.

The first International Conference on Information Processing, held in Paris, June 15–20, 1959, was the result of over two years of planning work. The original impetus came from I. L. Auerbach and the IRE-PGEC. The national Joint Computer Committee of the IRE, AIEE and ACM supported the effort, and a letter proposing such a conference was sent to UNESCO and to individuals and societies in other countries. The response was favorable, and a committee was formed by UNESCO to organize the conference. Out of these meetings there also grew the International Federation of Information Processing Societies, which will sponsor the next conference.

Over 1500 people from all over the world attended 11 planning sessions and 12 symposia. "Information Processing," the Proceedings of the Conference, contains the 60 papers presented, as well as summaries of the symposia presentations. The papers are mostly in English, although 6 are in French. All are abstracted in English, German, Russian and Spanish.

The papers are organized into the following chapters:

I. "Methods of Digital Computing" (mathematical applications)

II. "Common Symbolic Language for Computers" (automatic programming, especially ALGOL)

III. "Automatic Translation of Languages"

IV. "Pattern Recognition and Machine Learning"

V. "Logical Design of Computers"

VI. "Computer Techniques of the Future" (new components)

VII. "Miscellaneous Topics."

The papers are written for those sophisticated in the field, particularly for the hardware designers and scientific users. They present a good survey of the state of the art in 1958. The Proceedings is thus a reference work similar to the Proceedings of the East-

ern and Western Joint Computer Conferences, but more extensive than any of the latter. The quality of the papers is generally high. Workers in each specialty will probably find the material on their subject to be familiar, since the papers were all preprinted for distribution at the conference and since some were presented before in one form or another. Nevertheless, it is the kind of book that can be opened at random with a high probability that something of interest to those in the computer business will turn up.

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RECENT BOOKS

A Guide to U. S. Indexing and Abstracting Services. Prepared by the Science and Technology Division, Library of Congress, Washington, D.C., for the National Federation of Science Abstracting and Indexing Service under a grant from the National Science Foundation (NFAIS), 301 E. Capitol St., Washington 3, D. C. $2.00.

Jeenel, Joachim, *Programming for Digital Computers.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $12.00. Of interest to the advanced worker in the field of programming. Reviewed in the March, 1960, issue of IRE TRANSACTIONS ON ELECTRONIC COMPUTERS.

Liverhant, S. E., *Elementary Introduction to Nuclear Reactor Physics.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $9.75. Deals with the physical principles of neutron chain reactors.

Ordway, F. I., III, and Wakeford, R. C., *International Missile and Spacecraft Guide.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $25.00. Pictures and technical details of significant foreign and domestic rockets, missiles, and spacecraft are presented.

Sangren, Ward C., *Digital Computers and Nuclear Reactor Calculations.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $8.50. Acquaints the nuclear engineer and the scientist with the use of digital computers in the design of nuclear reactors.

Talley, David, *Basic Carrier Telephony.* John F. Rider Publisher, Inc., 116 W. 14 St., N. Y. 11, N. Y. $5.75. Deals with the art of transmitting multiple telephone and signaling signals over a single radio circuit, cable or wireline; and with the growth of the telephone industry and the necessity for using carrier telephony.

Tudbury, Chester A., *Basics of Induction Heating.* John F. Rider Publisher, Inc., 116 W. 14 St., N. Y. 11, N. Y. $8.90. Presents principles of induction heating art of commercial heating equipment.

Washtell, C. C. H., *An Introduction to Radiation Counters and Detectors.* Philosophical Library, Inc., 15 E. 40 St., N. Y. 16, N. Y. $7.50. An introduction in descriptive form to some of the nuclear radiation detectors, dealing with British practice and equipment.

Scanning the Transactions

Islands have been put to many uses by mankind (*e.g.*, Elba, Bikini, Bedloe's Island, Alcatraz), but none so unusual as the role proposed for Deception Island in the South Shetland Islands near the Antarctic Circle. For it has been suggested that Deception Island might make an excellent very-low-frequency antenna, to be used for generating whistlers artificially. Its latitude, 63°, places it in a part of the world where the rate of occurrence of natural whistlers has been found to be very great. But more interesting is its size and shape. Deception Island is 30 kilometers long, which is precisely a half-wavelength at 5 kc, a frequency at which natural whistlers are very strong. Moreover, its shape is that of a thin annulus surrounding a lagoon which has a single narrow entrance through the ring. This strongly suggests that the island, surrounded as it is by an excellent conductor (the sea), would constitute a naturally resonant, VLF slot antenna of high quality having a radiation pattern approximating that of a vertical monopole. From what is known of the geology of the island it appears that it would be a good insulator. This crucial point could be checked simply by measuring the driving point impedance of the island, a relatively easy matter. The island would be "fed" at its midpoint by means of transmission lines, the other ends of which would be terminated in the sea on either side, approximately 1 kilometer away. Thus, Deception Island seems likely to take all honors as the largest radio component on earth. (M. G. Morgan, "An island as a natural very-low frequency transmitting antenna," IRE TRANS. ON ANTENNAS AND PROPAGATION, September, 1960.)

International radio terminology underwent several significant changes during the International Radio Conference held in Geneva last year. While the "new" usages agreed on at Geneva are not unfamiliar to PROCEEDINGS readers, the fact that they now have the force and stature of officially recognized international standards makes them worth repeating here. The first change deals with frequency band designations. Because such expressions as "very high," "ultra high," and "super high" have little meaning when translated into other languages, it has been decided to designate frequency bands by the number of the exponent of 10 which expresses the upper frequency limit of the band in question. Thus, the VHF band, which extends from 3×10^7 to 3×10^8 cps, would be called Band 8. This system, while not in common usage in the U. S. is already recognized in IRE Standards and by other authoritative sources. Also adopted were the terms "gigacycle" and "teracycle" for 10^9 and 10^{12} cps. These terms are rapidly coming into wide usage. They were adopted as "preferred" terms by the IRE Editorial Board just two months ago. Still another change was the adoption of "Hertz" for cycle per second, leading to new frequency abbreviations such as kHz instead of kc. The Hertz has not as yet found wide acceptance in the U. S., but no doubt will begin to appear in the literature more frequently in the future. (W. H. Watkins, "Report on World Radio Conference, Geneva, Switzerland, 1959," IRE TRANS. ON BROADCASTING, August, 1960.)

Delay lines. Propagation delay plays a vital role in virtually every area of electronics, whether it be radar, traveling-wave tubes or ultra-high-speed computers. The ubiquity of this phenomenon is perhaps sufficient justification for repeating a poem credited to W. L. Everitt, which he used to recite in the classroom when introducing the concept of delay in transmission line theory. His immortal lines on delay went as follows:

There was a dachshund once so long
He hadn't any notion
How long it took to notify
His tail of an emotion.

And so it was that though his eyes
Were filled with woe and sadness
His little tail went wagging on
Because of previous gladness.

(B. R. Myers, "Education in the space age," IRE TRANS. ON EDUCATION, September, 1960.)

Radar has found an interesting new use as a surveying instrument for locating the most suitable transmission paths for short range army communications networks. The problem is that of rapidly choosing sites for mobile radio equipment which must operate over irregular terrain at ranges of 1 to 10 miles with antennas which for military reasons can be mounted no higher than 15 feet from the ground. It has been found that by locating a mobile surveillance radar at a proposed site, the terrain backscatter pattern as seen on the PPI would indicate the paths which offer the least obstruction to radio transmissions, thereby indicating the most suitable areas for locating other communication points in the network. (C. E. Sharp and R. E. Lacy, "Radio coverage-area survey-instrumentation research," IRE TRANS. ON VEHICULAR COMMUNICATIONS, August, 1960.)

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Antennas and Propagation	AP-8, No. 5	$1.50	$2.25	$4.50
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* Libraries and colleges may purchase copies at IRE Member rates.

Antennas and Propagation

VOL. AP-8, NO. 5, SEPTEMBER, 1960

Comparison of Experimental with Computed Tropospheric Refraction—W. L. Anderson, N. J. Beyers, and R. J. Rainey (p. 456)

Limits of applicability of ray tracing in computing tropospheric refraction at White Sands Missile Range have been further explored. 286 comparisons were made, all for a path from radar to fixed beacon of about 45 miles and an elevation angle of 17.99 milliradians. A horizontally stratified atmosphere was assumed. Refractive index profiles were prepared from a variety of weather data, and classified A, B, C, or R, in descending order of reliability prior to ray-tracing calculations.

Angle observations were made with an FPS-16 C-band radar having a quoted instrumental accuracy of 0.14 milliradian rms. Angle readings varied from 18.36 to 20.54 milliradians with mean of 19.01 milliradians and standard deviation of 0.41 milliradian.

The rms deviation of computed from experimental angles ranges from 0.28 to 0.41 milliradian for different classes of data. The ratio of this deviation to the deviation from over-all mean varies from 0.68 for Class A to 1.00 for Class R. Thus the improvement over a "standard atmosphere" varies from 32 per cent to 0, and correlates directly with quality of weather information.

For this experiment it is concluded that most of the rms elevation angle error is contributed by atmospheric conditions. Although ray tracing methods provide a significant correction when sufficiently good weather information is available, there still remains a large uncertainty not accounted for by equipment.

The Flared Slot: A Moderately Directive Flush-Mounted Broad-Band Antenna—J. W. Eberle, C. A. Levis, and D. McCoy (p. 461)

The flared slot is a simply constructed antenna designed to be flush-mounted in a horizontal surface. Its pattern is a moderately directive pencil beam over a 4:1 frequency band, and its impedance bandwidth is essentially that of a coaxial line-to-ridged waveguide transition. Both vertically and horizontally polarized models are discussed.

Mathematical and Experimental Studies of a Wide-Band Vertically Polarized Antenna—P. Foldes (p. 469)

Theory and experimental results are presented for an omnidirectional, vertically polarized antenna. One of the developed models has 7 db gain, VSWR <1.7, and 50 kw power handling capacity in a 2:1 frequency band. The new antenna is simple in construction and has a single coaxial feed point.

Maximally Flat and Quasi-Smooth Sector Beams—A. Ksienski (p. 476)

Methods for the synthesis of sector beams are presented and illustrated with examples. First a perfectly smooth and square beam, the equivalent of the maximally flat transfer function of network theory, is discussed and synthesized. It is then shown that realization of such beams by means of relatively short arrays is feasible when a certain amount of oscillation is permitted. The resulting beam approximates the maximally flat one in an almost equal ripple fashion, the amplitude of the ripple depending on the length of the array and the desired slope of the main beam.

A Geometrical Optics Method of Pattern Synthesis for Linear Arrays—Howard E. Shanks (p. 485)

A unique phase expression is given which

allows synthesis of a wide variety of shaped radiation patterns for a linear array. The accompanying amplitude distribution is found by a stationary phase evaluation of the radiation integral and is shown to be functionally related to the desired radiation pattern. Because of the optical type of approximation (stationary phase) used to evaluate the radiation integral, the method is most applicable to antennas with large size-to-wavelength ratios.

Gain Limitations of Large Antennas—Robert C. Hansen (p. 490)

Arrays of many elements and continuous aperture antennas are very often designed using Taylor's "approximate" space factor which is optimum in resolution, *i.e.*, it offers minimum beamwidth for a given sidelobe ratio. This space factor has a number of equal level sidelobes followed by tapered lobes, the latter being necessary to make the aperture function physically realizable. For conventional antennas, these designs offer gain which, while not that of a uniform aperture, is quite satisfactory. Very long (or large) antennas with their narrow beamwidths, however, may exhibit a severe gain degradation due to large energy contained in the sidelobes. Thus the sidelobe behavior must be carefully controlled to realize the full gain available. Data are presented for the gain variation of various Taylor equal sidelobe designs, and of the modified sin x/x one-parameter family as a function of aperture length.

Some Equivalences Between Equally and Unequally Spaced Arrays—Sheldon S. Sandler (p. 496)

General analytical expressions are presented for unequally spaced arrays. These relations allow for the analysis of the nonunitormly spaced array in terms of its equivalent uniformly spaced array. The inherent broad-band qualities of the nonuniformly spaced array are discussed. Some equivalence is made between the amplitude and spatial variation in the uniformly and nonuniformly spaced array. The general synthesis problem is discussed. An array with monotonically increasing interelement spacings is presented as an example of the theory.

The Efficiency of Launching Surface Waves on a Reactive Half Plane by an Arbitrary Antenna—Julius Kane (p. 500)

The following two-dimensional problem is considered: An arbitrary antenna is located at the origin which launches surface waves on a reactive half plane $y=0$, $x<0$. It is assumed that the other half plane, $y=0$, $x>0$, is occupied by a perfect conductor. The antenna is approximated by a distribution of multipoles easily found from the antenna's radiation pattern. With this approximation, simple algebraic expressions are obtained for the ratio of the power delivered to the surface wave as compared to the power in the radiated cylindrical field.

Surface-Wave Luneberg Lens Antennas—C. H. Walter (p. 508)

It is demonstrated that a surface-wave structure having isotropy in the plane of the structure can be operated as a lens. The index of refraction is given for a dielectric slab on a ground plane, a bed of metal posts on a ground plane and a parallel-plate structure with one plate filled with holes. The necessary radial variations in index for several Luneberg-type lenses are given. These include the conventional Luneberg lens, Gutman's modification, the rim-fed Luneberg adapted to a spherical surface and a modification adapted to a spherical surface and collimation of the rays at an arbitrary angle with respect to the plane on which the lens is mounted. An approximate analysis of surface-wave lenses based on optics is described.

Communications (p. 516)
Contributors (p. 531)

Broadcasting

Vol. BC-6, No. 2, August, 1960

Report on World Radio Conference, Geneva, Switzerland, 1959—William H. Watkins (p. 3)

A report on the world radio conference held in Geneva, Switzerland, during the last four months of 1959 is presented. The Conference was held under auspices of the International Telecommunication Union, the specialized agency of the United Nations which deals with radio and other telecommunication matters at international level. The organization of the conference is described, to show the mechanism by which engineers and telecommunication administrators from many nations work together to produce what amounts to international legislation in a highly technical field. The four principal work areas of the conference were: radio spectrum allocations; frequency management techniques; definition of technical terms and establishment of technical standards; and operating regulations in the aeronautical and maritime radio services. Major decisions of interest to engineers in the United States are summarized, including new terminology which engineers will be expected to learn gradually in the years ahead.

Future Possibilities for Film Room Mechanization—James H. Greenwood (p. 8)

New equipment is proposed for simplifying the telecasting of slides and film. All $2'' \times 2''$ slides currently in use are carried in one magazine. They are aired by momentary projection onto a storage tube which is then read out continuously while the slide is changed.

Instantaneous start of a film projector is accomplished by starting the parts having high inertia—motor and sound flywheel—before engaging the film; and running in this condition until show time.

A cartridge-loaded film projector for short segments carries sound on magnetic tape. Both tape and film are automatically cued to starting position.

Directional Antennas for Television Broadcasting—George H. Brown (p. 13)

Desirable operating characteristics and other requirements of directional transmitting antennas suitable for television broadcasting are considered. Basic limitations on directivity are imposed by long-distance propagation phenomena and reflections of signals into the weak-signal zones. Previous experience with directional antennas used for television broadcasting is reviewed. A number of appropriate antenna types are suggested.

Service Area of an Airborne Television Network (Abstract)—Martin T. Decker (p. 20)

Some Engineering Aspects of Video Tape Recording Production—Edward E. Benham (p. 21)

A Modern TV Transmitter Plant Input System—Joseph L. Stern (p. 25)

All the complex products of the television studio plant flow to the viewing audience through the transmitting plant and as the final link in the system it has an extremely heavy responsibility. To meet this responsibility the transmitting plant must have an input and control system that can act like a nerve center; be capable of continuous quality monitoring and instant corrective action. As part of a program of plant modernization and quality improvement a completely new and simple transmitting plant input system was designed. This paper outlines the study of transmitting plant functions and responsibilities that dictated the design, and describes the circuit and operating features of the resultant system.

A Special Effects Amplifier for Non-Composite or Composite, Monochrome or Color Television Signals—Ralph C. Kennedy (p. 39)

Remote Control of TV Microwave Equipment—John B. Bullock (p. 47)

Recent actions by the FCC have made it possible for privately owned and operated microwave links to be constructed by television stations. The need for remote control of the systems has brought about the development of some novel techniques for input switching, system reversal, and the transmission of control information. Hardware designed for these functions will be described.

The utility of a microwave relay system can often be greatly increased by the addition of such features as remote switching of system input, remote actuation of transmitter shutdown, fault location and reporting. If the system is one-way, the addition of remotely controllable reversing can add an entirely new dimension to its use.

The incorporation of these features into an equipment intended primarily for relaying TV signals poses certain unique problems. Such a system is generally one-way and has no microwave standby. Its entire base (video) band is occupied by the TV signal except for a limited space above the TV spectrum. Here one or more subcarriers may be added without degrading the TV signal, and these provide the only potential control channel(s) without adding parallel wire lines or radio equipment.

The RCA TVM-1A microwave relay equipment operates in the 5925–7425 Mc band. The transmitter utilizes a directly frequency-modulated klystron with one watt of RF power output. The receiver is a superheterodyne, employing a klystron local oscillator, crystal mixer and an IF centered at 130 Mc with a nominal bandwidth of 25 Mc. A repeater in a multihop system simply consists of a receiver and transmitter back-to-back. TV sound is included in the transmission by a subcarrier at 6.8 Mc, well above the video spectrum to avoid introduction of delay distortion by the combining filters. The equipment may be mounted in portable carrying cases or in standard 19-inch racks. Accessory units which make possible the remote control functions to be described are designed for integration in the rack mounted system only, inasmuch as such facilities are not usually needed in portable systems.

This paper will describe the methods worked out at RCA to provide simple and economical solutions to remote control, fault location and switching operations in the TVM-1A microwave relay equipment.

Broadcasting

Vol. BC-6, No. 3, September, 1960

On This Issue—The Editor (p. 1)
Comments Concerning FM Broadcast Standards of the Federal Communications Commission—O. W. B. Reed, Jr. (p. 2)
The Engineering Aspects of an Educational FM Network—Roger E. Peterson (p. 5)

FM radio relay techniques have been demonstrated since the early days of FM broadcasting. Several groups of commercial FM stations have formed small, loosely knit networks with varying degrees of technical performance and acceptance by the public. Educa-

tional institutions have been slow to realize tangible benefits from FM broadcasting.

The current resurgence of FM broadcasting in general and a new evaluation of the potential of aural broadcasting for education has prompted detailed plans for an extensive FM educational network. Educators recognize their need to know more; to exchange information and ideas with each other and with the faculties of other institutions; in brief to communicate. It is a special demand, sophisticated and vital to the educational system of today. It has not been met with commercial radio and television. The educational FM radio network described herein is designed to meet the needs of the educational system of today.

Multiplexing techniques would be used to enhance program handling capability and also provide remote control features allowing important economies in operation.

The Use of Phase Equalizers to Improve the Transient Response of a Television Transmitting System—J. K. MacDonald (p. 12)

A Description of the Tunnel Diode and its Applications—John W. Wentworth (p. 21)

Transmission of Colour Television Signals—J. Davidse (p. 31)

The paper discusses the transmission of color television signals according to the NTSC system. The choice of the chrominance signals, their bandwidths, and the subcarrier frequency is discussed. The consequences of the method of gamma correction and of deviations from the constant-luminance principle are considered. The significance of the statistics of the chrominance signal is pointed out.

Stereophonic Broadcasting Using Pulse-Amplitude Modulation—H. F. Mayer and F. Bath (p. 41)

For transmission of the two stereo signals of a stereophonic radio broadcast, a time multiplex method operating with pulse amplitude modulation is proposed. The transmitting principle is described. The questions of compatibility, transmitter excitation, baseband width and radio bandwidth, signal-to-noise ratio for single-channel and stereo reception, and synchronization are discussed. An account is given of the results obtained in trial workouts.

Education

Vol. E-3, No. 3, September, 1960

Editorial—Wilbur R. LePage (p. 65)

Education in the Space Age—Basil R. Myers (p. 66)

The dynamic persistence of scientific and technological development in the past two decades has aroused an intensity of public interest in education which is higher today than at any time in history. Perhaps the greatest impact of scientific progress has been the emergence of a more universal view of human society as a whole. We have two cultures—that of science, and that associated with the more traditional values of the humanities.

This paper represents a technologist's attempt to present a bird's eye view of the broad educational scene, both national and international. In discussing the major issue of the controversial scientist-vs-the-non-scientist problem, the author finds that it is the non-scientist—the humanist, the artist, the literary intellectual, the politician—who is the most unbalanced member of our society, in terms of the formal education which he receives.

A Laboratory-Problem Centered First Course for Electrical Engineers—E. J. Young and E. M. Williams (p. 78)

The objectives of Carnegie's first course for electrical engineering students are the conversion of basic science into "root content" and the development of professional method. Both objectives are met by a program of the solution of a series of professionally difficult but scientifically simple problems, which require both analytical and experimental work. All laboratory work is planned by the student, who takes major responsibility for its conduct and is permitted to make mistakes. Each new problem is used but once. Typical problem solutions require two to three weeks and two to three laboratory periods for completion. Recitation and laboratory work are integrated.

Several examples of problems are given, together with typical student plans, student reports, and instructor's comments. General rules for planning such a course are presented. The success of the course and its effect on the rest of the curriculum are discussed.

Circuit Theory in a Unified Curriculum—G. F. Paskusz and B. Bussell (p. 84)

Conventional circuit analysis courses are generally taught by electrical engineering instructors in a proprietary fashion. The underlying reasons for this state of affairs are history and utility; circuit theory was developed from a consideration of electric circuits and found its most important area of application there. Intrinsically, however, circuit theory is no more "electrical" than arithmetic is "financial."

This paper explains the basis for an abstract circuit course—"abstract" in the sense that the circuit connotation is not necessarily electrical but may, with equal validity, be mechanical, acoustical, thermal, etc. Generalized circuit variables are classed as either potential or flow variables, and generalized circuit parameters are then categorized according to their behavioral equations (Hooke's Law, Ohm's Law, Fourier's Law, etc.). A simple systematic method of finding a circuit diagram for a system composed of such parameters is shown.

Use of Educational Aids in Electronics—Harry E. Stockman (p. 89)

The Penultimate Teaching Machine—S. Seshu (p. 100)

This paper discusses the possibility of building an electronic transducer to take the place of the existing teaching machines. The tasks that can be performed by such a machine as well as the advantages and risks involved are discussed.

Contributors (p. 101)

Information Theory

IT-6, No. 4, September, 1960

On the Statistical Theory of Optimum Demodulation—J. B. Thomas and E. Wong (p. 420)

The multidimensional demodulation problem is considered from the point of view of statistical estimation theory and *a posteriori* most probable signal estimates are derived. Correlated signals and noises are treated. This formulation yields a set of two matrix integral equations which must be solved for the optimum estimates.

For amplitude modulation, the problem reduces to that of finding a set of time varying filters which are, again, solutions to a matrix integral equation. Special cases such as two-receiver systems, quadrature modulation, and single-sideband have particularly simple representations and are considered in some detail.

On the Mean-Square Noise Power of an Optimum Continuous Filter for Correlated Noise—Marvin Blum (p. 426)

This paper presents the equations for the mean-square error of the output of a continuous finite memory filter. The filter output error is unbiased for arbitrary input polynomials up to degree n, and has minimum variance. The input is taken as a polynomial of degree n plus random stationary noise. Noise processes are considered, 1) where the noise is exponentially correlated, and 2) in the white noise case. The solution for a desired output which is an arbitrary fixed linear operation on the input polynomial is given.

Tables and graphs of the mean-square error for the derivative and prediction operator for the 0th, 1st, and 2nd derivatives are presented, and for input polynomials up to the 6th degree.

Some Quantum Effects in Information Channels—T. E. Stern (p. 435)

In this paper the quantum nature of electromagnetic radiation is used as a basis for a mathematical model of a continuous channel. It is shown that this "photon channel" model leads to more realistic conclusions regarding information transmission. Among the results obtained are:

1) the maximum entropy for a narrow-band source under an average power limitation,
2) the frequency distribution (Bose-Einstein) for a wide-band power-limited source,
3) the transmission rate through a Poisson channel with additive Poisson noise.

Spectral Analysis of a Process of Randomly Delayed Pulses—M. V. Johns, Jr. (p. 440)

Formulas are obtained for the steady-state covariance function of a process of pulses separated by independent random time delays. The pulses considered may be either stochastic or deterministic in character. Since such processes approach stationarity in time their steady-state spectral density functions may be obtained. Explicit results are given for two examples.

Binary Codes with Specified Minimum Distance—Morris Plotkin (p. 445)

Two n-digit sequences, called "points," of binary digits are said to be at distance d if exactly d corresponding digits are unlike in the two sequences. The construction of sets of points, called codes, in which some specified minimum distance is maintained between pairs of points is of interest in the design of self-checking systems for computing with or transmitting binary digits, the minimum distance being the minimum number of digital errors required to produce an undetected error in the system output. Previous work in the field had established general upper bounds for the number of n-digit points in codes of minimum distance d with certain properties. This paper gives new results in the field in the form of theorems which permit systematic construction of codes for given n, d; for some n, d, the codes contain the greatest possible numbers of points.

On Decoding Linear Error-Correcting Codes—I—Neal Zierler (p. 450)

A technique is described for finding simply computable numerical-valued functions of a received binary word whose value indicates where errors in transmission have occurred. Although it seems that a certain condition must usually be fulfilled for such functions to exist, or for our method to constitute an efficient procedure for finding them, there is, on the one hand, a strong tendency for "good" codes to satisfy the condition, while, on the other, it appears to be straightforward to construct codes which are good for a specified channel and also fulfill the condition. An advantage of the resulting decoding procedure is that it corrects and detects all possible errors; more precisely, if a word u is received and the coset $\bar{u}$ to which u belongs has a unique leader e, the procedure concludes that $u+e$ was sent, while if u has no unique leader, that fact, along with the weight of $\bar{u}$ (and sometimes a little more) can be indicated. The ideas and tech-

niques are illustrated by the construction of decoding procedures for the perfect (23, 12) three-error-correcting code.

Encoding and Error-Correction Procedures for the Bose-Chaudhuri Codes—W. W. Peterson (p. 459)

Bose and Ray-Chaudhuri have recently described a class of binary codes which for arbitrary m and t are t-error correcting and have length 2^m-1 of which no more than mt digits are redundancy. This paper describes a simple error-correction procedure for these codes. Their cyclic structure is demonstrated and methods of exploiting it to implement the coding and correction procedure using shift registers are outlined. Closer bounds on the number of redundancy digits are derived.

Synchronization of Binary Messages—E. N. Gilbert (p. 470)

When messages are transmitted as blocks of binary digits, means of locating the beginnings of blocks are provided to keep the receiver in synchronism with the transmitter. Ordinarily, one uses a special synchronizing symbol (which is really a third kind of digit, neither 0 nor 1) for this purpose. The Morse code letter space and the teletype start and stop pulses are examples. If a special synchronizing digit is not available, its function may be served by a short sequence of binary digits P which is placed as a prefix to each block. The other digits must then be constrained to keep the sequence P from appearing within a block. If blocks of N digits (including the prefix P) are used, the prefix should be chosen to make large the number $G(N)$ of different blocks which satisfy the constraints. Lengthening the prefix decreases the number of "message digits" which remain in the block but also relaxes the constraints. Thus, for each N, there corresponds some optimum length of prefix.

For each prefix P, a generating function, a recurrence formula, and an asymptotic formula for large N are found for $G(N)$. Tables of $G(N)$ are given for all prefixes of four digits or fewer. Among all prefixes P of a given length A, the one for which $G(N)$ has the most rapid growth is $P=11\cdots1$. However, for this choice of P, the table of values of $G(N)$ starts with small values; $11\cdots1$ does not become the best A-digit prefix until N is very large. At these values of N, the $(A+1)$-digit prefix $11\cdots10$ is still better. The tables suggest that, for any N, a best prefix can always be found in the form $11\cdots10$, for suitable A. Taking $P=11\cdots10$ and $A=[\log_2(N\log_2 e)]$ it is shown that $G(N)$ is roughly $0.35N^{-1}2^N$. This result is near optimal since no choice of P can make $G(N)$ exceed $N^{-1}2^N$.

Analytic Inversion of a Class of Covariance Matrices—William A. Janos (p. 477)

The sample covariance matrix arising out of finite memory linear least squares estimation over a set of equally spaced time points, is inverted by spectral methods (operationally referred to as the z transform). It is shown that the complexity of the problem depends only upon the complexity of the input correlation function. The final solution is shown to reduce to the inversion of a triangular system of linear equations of an order less than half the degree of the denominator of the input power spectral density function.

An Isospectral Family of Random Processes—Richard A. Silverman (p. 485)

We construct a family of random step functions $\{x_n(t)\}$ whose members all have the same power spectrum and such that as $n\to\infty$, $x_n(t)$ converges to $x_\infty(t)$, the Gaussian process with the same spectrum. We illustrate the procedure for calculating the general multivariate distribution of the processes $\{x_n(t)\}$ by calculating the univariate, bivariate and trivariate distributions. We show how a suitably constructed univariate entropy can serve as an index of the extent to which $x_n(t)$ has approached the Gaussian limit $x_\infty(t)$.

On a Characterization of Processes for which Optimal Mean-Square Systems are of Specified Form—A. V. Balakrishnan (p. 490)

This paper presents the first results in an unconventional approach to the problem of mean-square optimization. Instead of obtaining a representation for the optimal operator for a process, we seek to characterize the class of processes for which the optimal operator is of specified form. If the processes are given, so that the multivariate characteristic functions are known, then our results can be used to tell whether it is possible for the optimal operator to have a specified form. The bulk of the paper pertains to the signal extraction problem where the signal and noise are independent and additive, and it is desired to estimate some function of the signal. Here, with a slight shift in viewpoint, we phrase the characterization problem in the following way: Given, for example, a noise process, determine the class of signal processes for which the optimal extraction system is of specified form. The case where the noise process is Gaussian comes in for special attention.

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Vehicular Communications

VC-9, No. 2, August, 1960

An Antenna for 30–50 Mc Service Having Substantial Freedom from Noise Caused by Precipitation Static and Corona—Ned A. Spencer (p. 2)

An antenna for service in the 30 50 Mc band is presented which affords a high degree of immunity from the noise of precipitation static and corona. Throughout this paper the noise caused by the impingement of charged particles is termed precipitation static. Corona is treated separately. A predominant feature of this design is a dielectric sheath enclosing the radiating conductor. A study of the causes of precipitation noise shows it to arise principally from a spark between a charged particle of precipitation and the antenna conductor. Noise from this source is estimated to be reduced about 30 db by the dielectric sheath. Experimental apparatus is described which enables quantitative measurements to be conducted in the laboratory. A standard approach is used to minimize corona effects.

The Past and Future Techniques of Vehicular Communications—E. W. Chapin (p. 6)

Radio Coverage-Area Survey-Instrumentation Research—C. E. Sharp and R. E. Lacy (p. 11)

The results of an experimental investigation on the problem of rapidly siting UHF and SHF radio relay type communications links in army grid networks are summarized. A requirement existed for a technique for locating suitable low loss transmission paths between fifteen foot high vehicular mounted antennas, separated at distances from one to ten miles in areas where irregular terrain obstructs propagation. Data are provided on an area survey instrumentation technique for this purpose employing radar principles.

Cryptographic Signaling Applied to Radio Communication Circuits—Owen E. Thompson (p. 17)

This paper considers the application of preprogramming to repetitious messages. After being programmed, transmission is by single-pulsed tone. The system provides more effective utilization of frequency, a high degree of security for message context, and a subscriber service much superior to that available in most present-day systems. After evaluation of the system described, consideration is given to the extension of these techniques to permit a high-speed, high-capacity, message service simultaneously with other services on a common channel.

Highway Alert Radio—E. A. Hanysz (p. 25)

A roadside-to-vehicle communications system, known as Hy-Com is proposed for bringing voice messages pertinent to driving conditions into moving vehicles. It fills the void left by visual aids which tend to lose their efficacy under conditions of darkness, inclement weather, heavy traffic and when driving at elevated speeds. Safer and more comfortable highway travel is indicated. Also, the improvement in communications can help to increase the volume of traffic handled by existing facilities.

The method proposed to transmit the message is through an amplitude-modulated magnetic induction field having a carrier frequency in the 10–14 kc range. The use of an induction field makes it possible to control the coverage to include only that traffic directly concerned. Interference to other services or other Hy-Com transmitters is avoided by the abruptness of the message field and thereby permits all Hy-Coms to operate on one and the same frequency channel.

The operation of Hy-Com is automatic once the ignition is turned on. It can share the audio output circuit and speaker with the standard broadcast radio in the car. The standard broadcast will be muted for the duration of the safety message and returned to normal after passing through the message field. If the radio is not on, the Hy-Com receiver will power the audio output stage so the message can be heard.

Portable transmitters using ferrite rod-type antennas can be used for emergency applications. Single wire, loop antennas, buried beneath the shoulder surface, and transmitters kept in tamper-proof concrete bunkers are proposed for permanent installations.

A New Colinear Antenna Array—A. H. Secord and W. V. Tilston (p. 36)

This paper describes the application of a phase changer which, when periodically distributed along a cylindrical conductor, controls the phase of the currents so that the radiation is concentrated in a plane perpendicular to the axis of the conductor. A theoretical analysis is made of a two-element structure, and the results are compared with those obtained by experiment. Further experimental results are given for multielement structures.

Can SSB Provide More Usable Channels in the Land Mobile Service?—A. V. Korolenko and N. H. Shepherd (p. 47)

Based on the assumptions made for system requirements and limited geographical separation of base stations, it has been shown that Sliver Band FM (SBFM $\pm$2.5 kc system deviation) can provide about the same degree of spectrum utilization as Single Sideband (SSB). Frequency spacing or required bandwidth is the main factor affecting spectrum utilization to be discussed. Required frequency spacing is controlled by equipment characteristics and geographic separation of stations. FM systems can be spaced closer together due to the modulation and noise characteristics of FM transmitters as compared to SSB transmitters.

Vehicular Communications in the City of Chicago Government—J. M. Veggeberg (p. 53)

A New Method of Providing Selective Calling and Noise-Immune Squelch for Radio Links—R. F. Herman (p. 57)

Rotary Wing Vehicular Communications—Richard A. Girvin (p. 62)

A Transistorized Receiver for 150-Mc Mobile Service—R. T. Myers (p. 70)

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

Acoustics and Audio Frequencies..... 1933
Antennas and Transmission Lines..... 1934
Automatic Computers..... 1935
Circuits and Circuit Elements..... 1935
General Physics..... 1937
Geophysical and Extraterrestrial Phenomena..... 1938
Location and Aids to Navigation..... 1940
Materials and Subsidiary Techniques.. 1940
Mathematics..... 1943
Measurements and Test Gear..... 1943
Other Applications of Radio and Electronics..... 1944
Propagation of Waves..... 1944
Reception..... 1945
Stations and Communication Systems.. 1945
Subsidiary Apparatus..... 1945
Television and Phototelegraphy..... 1945
Tubes and Thermionics..... 1946

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation. Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1., England.

ACOUSTICS AND AUDIO FREQUENCIES

534.154 3336

Piezoelectric Vibration Pick-Ups—W. Erler and A. Lenk. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 64–74; July, 1959.) The design and operating characteristics of various types of pick-up are reviewed.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears at the end of the Abstracts and References section.

The Index to the Abstracts and References published in the PROC. IRE from February, 1959 through January, 1960 is published by the PROC. IRE, June, 1960, Part II. It is also published by *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) and included in the April, 1960 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

534.213.4 3337

Theory of Wave Propagation in Irregular Waveguides—G. Ya. Lyubarskiĭ and A. Ya. Povzner. (*Zh. Tekh. Fiz.*, vol. 29, pp. 170–179; February, 1959.) Treatment of acoustic wave propagation in a waveguide of variable cross section by a generalization of the WKB method.

534.232-8:621.382.3 3338

Ultrasonic-Frequency Power Generator using Transistors—Y. M. Wei. (*Commun. and Electronics*, No. 46, pp. 1062–1065; January, 1960. The design of generators delivering 50–1000 W is described, and their performance is outlined.

534.26 3339

Bistatic Scattering from Totally Reflecting Flat Plates—P. W. Ankerman and R. A. Rubega. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 478–481; April, 1960.) An investigation of the characteristics of the scattering of sound from a flat plate of length 5 λ.

534.283 3340

Phonon Viscosity and its Effect on Acoustic Wave Attenuation and Dislocation Motion—W. P. Mason. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 458–472; April, 1960.) Measurements of the attenuation of sound in metals and nonconducting crystals are in agreement with the calculated values of dislocation damping by phonon viscosity.

534.4:551.578.1 3341

The Sound of Rain—M. R. Schroeder and B. F. Logan. (*Frequenz*, vol. 13, pp. 229–234; August, 1959. In English.) The statistics of the sound of rain falling on a plane surface are discussed, and the generation of such "rain noise" by electronic means is described, which produces a more realistic effect than the Gaussian noise usually employed for simulation purposes.

534.6:621.374 3342

Acoustic Electronic Pulse Equipment—E. Skudrzyk. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 565–571; May, 1960.) Details of the construction and operation of a pulse generator, amplifier and transducer are given. The frequency range of the generator is 50 cps-100 Mc, and the pulse repetition rate can be varied from 0.1/second to 25000/second with pulse duration $\frac{1}{3}$ μsec-100 msec.

534.6.087.9 3343

A Semi-automatic Analyzer for Reverberation Measurements—A. Lauber and K. Hess. (*Tech. Mitt. PTT.*, vol. 37, pp. 225–230; June 1, 1959. In German and French.) The equipment described facilitates the evaluation of reverberation curves recorded by hypsograph.

534.612.2 3344

Microphone Diaphragm Null Method for Sound Pressure Measurement—W. Koidan. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 505–507; April, 1960.) Sound pressure is determined from a measurement of the driving voltage necessary to keep the diaphragm of a condenser microphone from moving when it is exposed to a sound field. The method is applicable when it is desired that the microphone should present an infinite impedance to the medium and absorb no energy from the sound field.

534.612.2.089.6 3345

Calibration for Carrier-Operated Microphones and other Reversible Transducers—M. D. Burkhard, E. L. R. Corliss, W. Koidan, and F. Biagi. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 501–504; April, 1960.) The modulated carrier system of Zaalberg van Zelst (9 of 1949) has been adapted for the measurement of sound pressures at low-noise levels by an insertion technique.

534.7 3346

The Excitation of the Ear by Continuous Noise and by Short Pulses—R. Oetinger. (*Nachrichtentech. Z.*, vol. 12, pp. 391–399; August, 1959.) The excitation of Corti's organ is investigated on the basis of subjective tests.

534.75 3347

The Monaural Directional Characteristic of the Human Ear—G. Jahn and S. Vogelsang. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 50–56; July, 1959.) Report on objective measurements of sound level at the membrane and comparison with the results of earlier subjective measurements.

534.78 3348

Speech Bandwidth Compression through Spectrum Selection—K. D. Kryter. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 547–556; May, 1960.) Intelligibility tests were made using *a*) single band-pass filters of different bandwidths and *b*) various configurations of one, two, and three band-pass filters of bandwidth 550 cps. Results showed that for equal performance the former system requires an effective bandwidth twice that of the best three-band system.

534.79 3349

A Graphical Method for the Determination of Sound Intensity and Loudness from Level Diagrams covering an Interval of a Third—E. Zwicker. (*Frequenz*, vol. 13, pp. 234–238;

August, 1959.) A template method is described, and template diagrams are given for evaluating loudness on the basis of sound level measurements based on intervals of one third. See also 3544 of 1959.

534.79 3350
The Threshold of Pressure on the Ear for Different Acoustic Phenomena—E. Zwicker. (*Frequenz*, vol. 13, pp. 238–242; August, 1959.) The threshold is defined as that level of sound pressure which, after being applied to the ear for one minute just gives rise to a sensation of pressure. This effect is investigated for pure tones as a function of frequency in the range 5–80 cps, for octave-band noise, and for pulse trains as a function of pulse duration.

534.79 3351
The Loudness of Tone Pulses of Differing Duration—E. Port. (*Frequenz*, vol. 13, pp. 242–245; August, 1959.) The results of a series of subjective tests are given, which indicate the difficulty of defining the time characteristics of the ear by a single-valued time constant.

534.845 3352
The Sound Absorption of Single Resonators with Omnidirectional Incidence of Sound and for a Linear Arrangement, in the Centre of a Room, on the Wall, along the Edge or in the Corner of a Room—W. Wöhle. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 56–61; July, 1959.) Results of calculations on interaction effects are given and discussed. See also 2593 of August.

621.395.613.38 3353
Speech Transmission by means of the Ear Microphone—J. Naujoks. (*Nachrichtentech Z.*, vol. 12, pp. 400–402; August, 1959.) Measurements of the transmission characteristics of ear-insert microphones are analyzed, and the results of intelligibility tests made with this type of microphone, used on a single channel and stereophonically, and with a throat microphone and a pressure-gradient microphone are compared. See also 2661 of 1957 (Black).

621.395.616.089 3354
Method for Measurement of $|E'/I'|$ in the Reciprocity Calibration of Condenser Microphones—W. Koidan. (*J. Acoust. Soc. Amer.*, vol. 32, p. 611; May, 1960.) A method is described by which the adjustment of an attenuator and a standard resistance give a direct reading of the ratio of the open-circuit voltage of the receiver to the source driving current.

ANTENNAS AND TRANSMISSION LINES

621.372.2 3355
A Graphical Method for Describing the Transformation Characteristics of Lossy Lines—K. E. Müller. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 61–64; July, 1959.) The application of the Smith diagram to lossy lines is described.

621.372.8:537.56 3356
The Backward Wave in a Metal Waveguide Filled with Longitudinally Magnetized Plasma—W. O. Schumann. (*Z. angew. Phys.*, vol. 11, pp. 333–335; September, 1959.) Wave propagation in a cylindrical waveguide is considered on the basis of an earlier analysis (1881 of 1957).

621.372.8:621.317.3 3357
Special Problems of Microwave Measuring Techniques using Waveguides—W. Krausse. (*Nachr. Tech.*, vol. 9, pp. 351–355; August, 1959.) Problems arising from dimensional and assembly tolerances; in particular, the reflections at points of misalignment are considered.

621.372.826 3358
Surface Wave Excitation and Propagation—J. B. Keller and F. C. Karal, Jr. (*J. Appl. Phys.*, vol. 31, pp. 1039–1046; June, 1960.) A geometrical theory based on the concept of a complex or imaginary ray is developed for the analysis of excitation and propagation of surface waves of known propagation constant.

621.372.826 3359
The Propagation of Surface Electromagnetic Waves along a Many-Wire System—N. A. Armand. (*Zh. tekh. Fiz.*, vol. 29, pp. 107–119; January, 1959.) Simplified boundary conditions are introduced and the dispersion equation is analyzed. A detailed treatment is given for the case of a two-wire line with and without dielectric coating. Two types of surface wave which are propagated with different phase velocity and different attenuation are shown to exist.

621.372.829 3360
On the Theory of Propagation of the H_{01} Wave in a Helix Waveguide—V. V. Malin and A. N. Sivov. (*Radiotekhnika i Elektronika*, vol. 4, pp. 433–439; March, 1959.) Estimation of losses of the H_{01} Wave in a dielectric located in the proximity of a helix waveguide made of a coiled rectangular or circular wire. The effect of the exterior dielectric envelope on the radiation losses is examined, and the additional attenuation produced by a slight ellipticity of the waveguide is discussed.

621.372.829 3361
Attenuation of H_{on} Waves in a Helix Waveguide—B. Z. Katsenelenbaum. (*Radiotekhnika i Elektronika*, vol. 4, pp. 428–432; March, 1959.) Expressions are derived for the losses caused by radiation and to ohmic resistance, and in the dielectric adjoining the surface of the waveguide.

621.372.852.323 3362
New Isolators for Radio Link Systems—J. Deutsch, W. Haken, and C. v. Haza-Radlitz. (*Nachrichtentech. Z.*, vol. 12, pp. 367–370; July, 1959.) Resonance-type ferrite-strip isolators for 4-, 6- and 7-kMc are described.

621.396.67:621.318.57 3363
A Circular-Polarization Duplexer for Millimeter Waves—R. G. Fellers. (*Commun. and Electronics*, pp. 934–937; January, 1960.) The use of closely spaced metal slot gratings in a duplex system with negligible power loss is described.

621.396.673.029.53 3364
Antenna Design for Maximum L. F. Radiation—G. J. Monser and W. D. Sabin. (*Electronics*, vol. 33, pp. 84–86; June 3, 1960.) Top-loading techniques for vertical antennas acting as half-wave elements at 1–2 Mc are described. Data obtained on test antennas showing efficiencies up to 90 per cent are given.

621.396.673.095 3365
The Electric Field at the Ground Plane near a Top-Loaded Monopole Antenna with Special Regard to Electrically Small L- and T-Antennas—H. L. Knudsen and T. Larsen. (*J. Res. NBS*, vol. 64D, pp. 139–151; March/April, 1960.) A calculation of the field for antennas having a known current distribution.

621.396.674.1 3366
Radiation Fields of Circular Loop Antennas by a Direct Integration Process—E. J. Martin, Jr. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 105–107; January, 1960.) Generalized expressions are derived which describe the field at any point in space for loops with standing-wave or traveling-wave current distributions.

621.396.676 3367
Transmission-Line Missile Antennas—R. King, C. W. Harrison, Jr., and D. H. Denton, Jr., (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 88–90; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

621.396.677.091.22 3368
Calculation of the Gain of an Aerial with an Elliptical Radiation Pattern—I. S. Khevrunin. (*Radiotekhnika i Elektronika*, vol. 4, pp. 381–383; March, 1959.) Formulas are derived for calculating the antenna gain and polar diagram which, according to the direction of the field potential, can be represented by triaxial ellipsoids with one of the axes orientated in the direction of maximum radiation.

621.396.677.3 3369
Broadband Log-Periodic Antennas—R. L. Bell. (*Electronics*, vol. 33, pp. 58–59; June, 17 1960.) A short review is given, with examples of typical arrays and their directional characteristics.

621.396.677.3 3370
Back-Scattering Cross-Sections of Cylindrical Wires of Finite Conductivity—E. S. Cassedy and J. Fainberg. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 1–7; January, 1960.) Abstract, PROC. IRE, vol. 48, pt. 1, p. 1189; June, 1960.)

621.396.677.3 3371
A Highly Directive Rotating Array for 16 Mc/s—J. A. Thomas and R. W. E. McNicol. (*Nature (London)*, vol. 187, pp. 398–399; July 30, 1960.) A horizontal array for ionospheric back scatter sounding is described. It consists of four three-element Yagi antennas separated by 1.1 λ, mounted on a truss 228 feet long, pivoted at its center and supported by four bogies. The effective beamwidth is 8°, and the sidelobe attenuation at least 19 db. Other characteristics are given.

621.396.677.71 3372
A Slot with Variable Coupling and its Application to a Linear Array—R. Tang. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 97–101; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

621.396.677.8 3373
Transient Behaviour of Aperture Antennas—C. Polk. (PROC. IRE, vol. 48, pp. 1281–1288; July, 1960.) It is shown that the steady-state main lobe is established within $\tau/2+R/c$ second where τ is the carrier period and R the distance of the point of evaluation from the center of the aperture.

621.396.677.81 3374
On Uniform and Linearly Tapered Long Yagi Antennas—D. L. Sengupta. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 11–17; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1189; June, 1960.)

621.396.677.83 3375
The Calculation of Reflector-Antenna Polarized Radiation—L. E. Raburn. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 43–49; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, pp. 1189–1190; June, 1960.)

621.396.677.83 3376
Design of Circular Apertures for Narrow Beam Width and Low Side Lobes—T. T. Taylor. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 17–22; January, 1960.) An extension of an earlier analysis (984 of 1956). For tables of the functions, see *ibid.*, pp. 23–26 (Hansen).

621.396.677.832 3377
Impedance of a Corner-Reflector Antenna as a Function of the Diameter and Length of

the Driven Element—A. C. Wilson. (*J. Res. NBS*, vol. 64D, pp. 135–137; March/April, 1960.) The measurements were made on a monopole in a corner-reflector over an image plane, and the results are given as a family of curves.

621.396.677.833.012.12 3378
Radiation Pattern Synthesis with Sources Located on a Conical Surface—A. Ishimaru and G. Held. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 91–96; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

621.396.677.85 3379
Calculation of Phase Velocities of Radio Waves in an Artificial Metal-Dielectric—Ya. N. Fel'd and L. S. Benenson. (*Radiotekhnika i Elektronika*, vol. 4, pp. 417–427; March, 1959.) A characteristic equation for the phase velocity is derived, the solution of which is stationary in relation to the variation of current distribution on the elements of the structure.

621.396.677.85:537.226 3380
Scattering by an Infinite Array of Thin Dielectric Sheets—R. E. Collin. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 62–67; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

AUTOMATIC COMPUTERS

681.142 3381
Electronic Computers 1958/59—H. K. Schuff. (*VDI Z.*, vol. 102, pp. 701–705; June 11, 1960.) Annual review of international developments in digital- and analogue-computer design and installations. 76 references.

681.142 3382
Physical versus Logical Coupling in Memory Systems—J. A. Swanson. (*IBM J. Res. & Dev.*, vol. 4, pp. 305–310; July, 1960.) An evaluation is made of the optimum size of a cell for maximum information storage.

681.142:621.318.57 3383
A Small High-Speed Transistor and Ferrite-Core Memory System—W. L. Shafer, Jr., W. N. Toy and H. F. Priebe, Jr. (*Commun. and Electronics*, pp. 763–769; January, 1960.) A description is given of the design and operation of a 400-bit storage system used in a time-division switching system where the number of circuits controlled required a simultaneous read-out of 20 bits.

681.142:621.318.57:537.312.62 3384
The "Persistor"—a Superconducting Memory Element—E. C. Crittenden, Jr., J. N. Cooper, and F. W. Schmidlin. (PROC. IRE, vol. 48, pp. 1233–1246; July, 1960.) The persistor consists of a superconducting inductor in parallel with a normally superconducting element. A current pulse causes a current to be established in the persistor; another similar pulse causes no change, but one of opposite sign reverses the current. Pulses of 15-mμsec duration can be used at a repetition rate of 15 Mc, and the elements are well suited to printed circuit production.

681.142:621.318.57:621.382.23 3385
Tunnel-Diode Logic Circuits—W. F. Chow. (*Electronics*, vol. 33, pp. 103–107; June 24, 1960.) Equations which relate diode tolerances to those of other circuit components for analogue-threshold logic and majority logic use are derived. Practical tunnel-diode logic circuits are believed to be possible.

681.142:621.318.57:621.385.6 3386
A New Class of Switching Devices and Logic Elements—P. R. McIsaac and I. Itzkan. (PROC. IRE, vol. 48, pp. 1264–1271; July, 1960.) With switching devices employing microwave-tube elements, switching at 1-mμsec intervals becomes possible. An experimental traveling-wave interaction-type tube has been built to demonstrate the ability of one microwave signal to control another.

68.142:621.372.44 3387
Parametron Computer Circuits—K. Nagamori. (*Electronics*, vol. 33, pp. 73–78; June 3, 1960.) The characteristics of parametrons and their applications are discussed. See 3588 of 1959 (Goto).

681.142:621.374.5 3388
Ripple-Type Time-Delay Networks using Elliptic Functions—J. R. Kiseda and D. J. Ford. (*Commun. and Electronics*, No. 46, pp. 996–1002; January, 1960.) The design of simulation networks in the frequency domain is presented.

681.142:621.375.3 3389
Magnetic-Amplifier Binary-to-Analogue Conversion—I. Danylchuk and D. Katz. (*Commun. and Electronics*, No. 46, pp. 909–912; January, 1960.) Three saturable reactors are used to convert a 10-digit input to an analog voltage driving a servo magnetic amplifier.

681.142:781.6 3390
Composing with Electronic Computers—P. G. Neumann and H. Schappert. (*Nachrichtentech. Z.*, vol. 12, pp. 403–407; August, 1959.) Two basic methods of musical composition by computer are discussed to illustrate the simulation of human thought processes.

CIRCUITS AND CIRCUIT ELEMENTS

621.316.5:518.4 3391
Synthesis of Switching Functions by Linear Graph Theory—W. Mayeda. (*IBM J. Res. & Dev.*, vol. 4, pp. 321–328; July, 1960.)

621.316.825 3392
Survey of Thermistor Characteristics—J. Van Dover and N. F. Bechtold. (*Electronics*, vol. 33, pp. 58, 60; July 1, 1960.) Characteristics and applications are tabulated for different constructions of the two basic types of thermistor.

621.316.84/.86 3393
The Long-Term Stability of Fixed Resistors—H. F. Church. (*Proc. IEE*, vol. 107, pt. B, pp. 377–385; July, 1960.) An investigation of the causes of long-term failure under practical conditions of use or storage. Test methods are critically discussed.

621.318.435 3394
A Mathematical Model for a Magnetic-Amplifier Reactor Core—H. W. Woodson. (*Commun. and Electronics*, No. 46, pp. 872–878; January, 1960. Discussion.) From domain theory, a model for the prediction of core behavior is proposed.

621.318.435:621.318.57 3395
Two-Step Switching of Saturable Reactors—B. W. Lovell. (*Commun. and Electronics*, No. 46, pp. 895–900; January, 1960.) The performance of the mathematical model of Woodson (3394 above) when subjected to abrupt flux reversals is compared with experimental data for Orthonol cores.

621.318.318.435:621.373.443 3396
D.C.-Charged Magnetic Pulse Modulator—E. W. Manteuffel and R. E. Cooper. (*Commun. and Electronics*, No. 46, pp. 843–850; January, 1960.) Design details and performance results are quoted for an experimental modulator consisting of one charging and two pulse-modulating stages, capable of delivering a peak power output of 110 kW, with a pulse width of 1.5 μsec, into a resistive load.

621.318.57:537.312.62:681.142 3397
The 'Persistor'—a Superconducting Memory Element—Crittenden, Cooper, and Schmidlin. (See 3384.)

621.319.4 3398
Some Mechanisms of Failure of Capacitors with Mica Dielectrics—A. A. New. (*Proc. IEE*, vol. 107, pt. B, pp. 337–364; July, 1960.) An analysis of failures, with a description of some methods of examination and dissection giving minimum loss of evidence.

621.372.5/.6 3399
The Representation of Multipoles by Scattering Matrix—E. Schuon and H. Wolf. (*Nachrichtentech. Z.*, vol. 12, pp. 361–336 and 408–415; July–August, 1959.) The scattering matrix is defined, and its properties are summarized. The application of the matrix to the calculation of multipole networks, including quadripoles, is detailed with tables giving the matrices of some basic circuits.

621.372.57:681.142 3400
Differentiating and Integrating Circuits with Positive Operational Feedback—A. I. Petrenko. (*J. Electronics Control*, vol. 8, pp. 59–66; January, 1960.) A general method of analysis and design is presented, and illustrated with a particular circuit.

621.372.57:681.142 3401
Valve Circuits with Normalized Transfer Function for Distortionless Electrical Differentiation—K. H. Kerber. (*Frequenz*, vol. 13, pp. 203–210; July, 1959. Correction, vol. 13, p. 306; September, 1959.) The circuit parameters and operating characteristics of several differentiating circuits for obtaining first- and second-order derivatives are given.

621.372.632:621.382.3 3402
An Equivalent Circuit for Transistor Mixers—H. Beneking. (*Arch. elekt. Übertragung*, vol. 13, pp. 313–319; July, 1959.) An equivalent circuit is derived for the earthed-emitter configuration, based on Giacoletto's circuit [see, *e.g.*, 1427 of April (Moortgat-Pick)]. Experimental confirmation of the validity of this theory is obtained.

621.373 3403
Oscillators and their Modes of Oscillation—W. Herzog. (*Nachrichtentech. Z.*, vol. 12, pp. 421–430; August, 1959.) A classification of oscillators in three types is proposed, based on the parameters of the passive frequency-determining quadripole which, together with an active quadripole, constitutes the oscillator. Several basic circuits are analysed in this way and the *Q*-factor of their modes of oscillation is evaluated.

621.373.029.6:621.317.733 3404
A Microwave Meacham Bridge Oscillator—W. R. Sooy, F. L. Vernon, and J. Munushian. (PROC. IRE, vol. 48, pp. 1297–1306; July, 1960.) A high-stability microwave oscillator has been constructed, consisting of a traveling-wave amplifier, cavity, microwave bridge, and bolometer. Experimental results are compared with theory.

621.373.4 3405
Quantitative Analysis of a Nonlinear Second-Order Differential Equation of an Self-Oscillating System with a Limited Increment Region—A. A. Luchina. (*Radiotekhnika i Elektronika*, vol. 4, pp. 440–448; March, 1959.)

621.373.431.1 3406
Stability Problems in Monostable Multivibrators—G. Kosel. (*Elektronische Rundschau*, vol. 13, pp. 277–279; August, 1959.) The influence of the amplitude of trigger pulses on multivibrator operation, and the dependence of pulse width on the trigger frequency are discussed. See also 2648 of August.

621.373.431.2 3407
A Graphical Analysis of the Blocking Oscillator—D. Chambers. (*Electronic Engrg.*, vol. 32, pp. 275–279, 348–354 and 430–434; May–July, 1960.) A detailed analysis of the circuit and its mode of operation is given, and a practical design procedure is described.

621.373.44:621.382.3 3408
Transistor Monostable Circuits—T. H. Brown and W. L. Stephenson. (*Mullard Tech. Commun.*, vol. 5, pp. 89–96; April, 1960.) Cross-coupled and emitter-coupled circuits are analysed to determine the circuit characteristics. Results obtained with two practical circuits show good agreement with theory.

621.374 3409
A Passive Waveform Shaping Circuit—M. J. Wright. (*Electronic Engrg.*, vol. 32, pp. 341–343; June, 1960.) An inductance and suitable semiconductor diodes are combined. Applications discussed include (*a*) 90° change in phase of a square wave; (*b*) mark-to-space ratio modulation of a square wave; (*c*) pulse generation, lengthening, and delay.

621.374 3410
A Simple Circuit for Producing a Voltage Proportional to the Width of a Repetitive Pulse—H. C. Bertoya. (*Electronic Engrg*, vol. 32, pp. 442–443; July, 1960.) Details of a circuit for deriving a direct voltage from a train of 25 volts 800-cps pulses of width proportional to the flow rate of a liquid are given. The voltage drives an integrating motor via a push-pull cathode follower.

621.374:621.382.333 3411
Complementary Transistor Circuits—F. Oakes and C. Thompson. (*Electronic Engrg.*, vol. 32, pp. 438–442; July, 1960.) By the use of *p-n-p* and *n-p-n* transistors connected back to back, a series of circuits become available for which there is no tube equivalent. They are symmetrical in operation, but do not necessarily require balanced transistors.

621.374.4:621.317.755 3412
Derivation of a Frequency Multiple of a Variable Frequency—V. Gavreau and A. Calaora. (*Compt. rend. acad. sci., Paris*, vol. 249, pp. 2026–2028; November 16, 1959.) A mask is placed across the face of a CRO displaying the variable frequency f. The mask outline is a sinusoid representing the frequency multiple n. It is viewed by a photocell whose output varies at a frequency nf. Equations defining the mask outline for any value of n are given.

621.374.5 3413
A Pulse Delay System—A. Karaminkov. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 42–49; July, 1959.) Details are given of a phase-shift circuit followed by a pulse-shaping stage; a maximum delay of about 1 msec can be achieved.

621.375.2.121 3414
Stagger-Tuned and Bandpass-Coupled Circuits in Wide-Band I.F. Valve Amplifiers—J. S. Singleton. (*Mullard Tech. Commun.*, vol. 5, pp. 48–59; February, 1960,) Formulas are derived for the replacement of any two equally detuned circuits in a stagger-tuned amplifier by a single band-pass coupled pair.

621.375.227 3415
A New Asymmetric Push-Pull Amplifier with Extremely Low Internal Impedance—H. Völz. (*Elektronische Rundschau*, vol. 13, pp. 242–244; July, 1959.) An output amplifier circuit is described in which the phase splitting triode is incorporated in one of the push-pull arms.

621.375.234 3416
Contribution on the Amplifier with Extremely Low Internal Impedance—H. Völz. (*Frequenz*, vol. 13, pp. 212–222; July, 1959.) Detailed investigation of the operation of a two-stage tube amplifier with combined negative voltage feedback and positive current feedback.

621.375.3 3417
Elimination of Coupling Problems in Multistage Magnetic Amplifiers—W. McMurray. (*Commun. and Electronics*, pp. 858–863; January, 1960.) Operational principles and design criteria are considered for two-stage half-wave and full-wave amplifiers with synchronous-switch coupling.

621.375.3 3418
Theory of 3-Phase Bridge Magnetic Amplifier: Part 1—Steady State—H. F. Storm and C. W. Flairty. (*Commun. and Electronics*, pp. 878–895; January, 1960.) A discussion of operation at minimum load with the control circuit, either open or closed, is followed by a study of conditions under normal excitation from minimum to half-load (saturation angles from 120° to 60° for constant load resistance), and from half-load to full-load (saturation angles 60° to 0°).

621.375.3:681.142 3419
Miniature High-Frequency Magnetic Amplifiers for Magnetic Memory Sensing—G. R. Briggs and K. C. Hu. (*Commun. and Electronics*, pp. 863–872; January, 1960.) A report of measurements on ferrite and metal tape cores with single-turn control windings and asymmetrical excitation at frequencies in the 100 kc-4 Mc range indicates stable amplification of signals as small as 10^{-10}W with response times of 10 μsec.

621.375.4:621.395 3420
Transistors in Audio- and Carrier-Frequency Amplifiers—K. Lamont. (*Electronic Tech.*, vol. 37, pp. 292–299; August, 1960.) The characteristics of common-emitter transistor amplifiers in relation to requirements for communication systems are discussed, and a design for a 25-db amplifier with 30-db feedback suitable for a five-channel telephone system is given.

621.375.9:538.569.4 3421
An Introduction to the Theory of Solid-State Masers—P. N. Butcher. (*Proc. IEE*, vol. 107, pt. B, pp. 341–351; July, 1960. Discussion, pp. 352–353.) Basic theory is summarized and used to evaluate the engineering characteristics of a traveling-wave maser employing the comb type of slow-wave guide.

621.375.9:538.569.4 3422
Nuclear-Magnetic-Resonance Maser—Benoit. (See 3463.)

621.375.9:538.569.4:537.311.33 3423
Quantum-Mechanical Semiconductor Generators and Amplifiers of Electromagnetic Oscillations—N. G. Basov, B. M. Vul, and Yu. M. Popov. (*Zh. eksp. teor. Fiz.*, vol. 37, pp. 587–588; August, 1959.) A note on the possibility of using the electron transitions between the conduction (or valence) band and the donor (or acceptor) impurity levels to obtain maser action.

621.375.9:621.372.44 3424
Theory of Single-Resonance Parametric Amplifiers—S. T. Fisher. (Proc. IRE, vol. 48, pp. 1227–1232; July, 1960.) A parametric amplifier that supports a resonance at the idler frequency only is capable of unlimited gain when operated either as a straight-through amplifier or as a lower-sideband up-converter. This makes a continuously tunable parametric receiver feasible.

621.375.9:621.372.44 3425
Unidirectional Paramagnetic Amplifier Design—M. W. P. Strandberg. (Proc. IRE, vol. 48, pp. 1307–1320; July, 1960.) The theory and design of a cavity paramagnetic amplifier are given. The Faraday effect is used to obtain unidirectional amplification, which can result in noise figures an order of magnitude better than other systems.

621.375.9:621.372.44:621.382.2 3426
Theory of the Diode-Reactance Amplifier with Parallel Circuits—W. Dahlke, R. Maurer, and J. Schubert. (*Arch. elekt. Übertragung*, vol. 13, pp. 321–340; August, 1959.) Detailed treatment of the theory and design of parametric amplifiers consisting of a variable-reactance semiconductor diode coupling two resonant circuits. Minimum noise conditions are evaluated, and design parameters are tabulated for three basic types of circuit. 35 references.

621.375.9:621.372.44:621.382.2 3427
Low-Noise Parametric Amplifier—R. C. Knechtli and R. D. Weglein. (Proc. IRE, vol. 48, pp. 1218–1226; July, 1960.) The noise temperature and pump power using a semiconductor diode can be predicted accurately from a measurement of the diode and circuit parameters; the ultimate lower noise limit depends on the product of diode cut-off frequency and the normalized capacitance swing.

621.375.9:621.372.44:621.385.63 3428
Space-Charge-Wave Parametric Amplifiers—Cook, Louisell, and Quate. (See 3688.)

621.375.9:621.382.23 3429
Diode Capacitors for Parametric Amplification—R. C. Knechtli, R. D. Weglein, and K. E. Mortenson. (*J. Appl. Phys.*, vol. 31, pp. 1134–1135; June, 1960.) Comment on 441 of February and author's reply.

621.376.22 3430
Limitations in Squaring with Simple Valve Circuits of Wide Bandwidth—R. M. Huey and J. E. Longfoot. (*Electronic Engrg.*, vol. 32, pp. 425–428; July, 1960.) The errors to be expected from squaring circuits (*a*) based on curvature of a tube characteristic with input applied to one electrode, and (*b*) using separate control electrodes with antiphase inputs are compared. Measurements have been made on the latter type.

621.376.3 3431
Network Response to Transient Frequency-Modulation Inputs—T. T. N. Bucher. (*Commun. and Electronics*, pp. 1017–1022; January, 1960.) An approximate method is developed for calculating the response when the input modulation varies rapidly during short periods, *e.g.*, when television or pulse signals are used for frequency modulation.

621.376.33 3432
Modulation Distortion in Frequency Modulation by Special Frequency-Band Limitation—E. Henze. (*Arch. elekt. Übertragung*, vol. 13, pp. 348–355; August, 1959.) The distortion factors are calculated which arise in two-, three- or four-double-tuned, critically coupled band filters used in FM receivers, under the assumption of linearized phase-characteristics and for

two different selectivity specifications. The amplitudes of combination tones are calculated for the case of three filters when two FM signals are applied.

GENERAL PHYSICS

535.215:535.34 3433

Light Waves in Crystals in the Region of Exciton Absorption and the Impurity Photoeffect—I. M. Dykman and S. I. Pekar. (*Zh. eksp. teor. Fiz.*, vol. 37, pp. 510–521; August, 1959.) In the frequency range for which the refractive index is much less than unity, the amplitudes of normal and longitudinal waves in a cubic crystal greatly exceed the amplitude of the incident wave. The photoionization of impurities is therefore much more intense. This explains the sharp maximum in the internal and external photoeffects observed experimentally in the frequency range corresponding to exciton absorption. See 409 of 1959 and 3058 of 1958 (Pekar).

537.222:537.533.9 3434

Experimental Investigation of Charge formation on a Dielectric Surface under the Influence of Electron Bombardment: Part 2—N. M. Karnaukhova and V. Ya. Upatov. (*Radiotekhnika i Elektronika*, vol. 4, pp. 521–526; March, 1959.) The kinetics of the formation of an actual charged spot produced on the surface of a dielectric by an electron beam beam have been investigated. The instrument and method are described, and it is shown that the conditions for the formation of positive and negative spots are different. For Part 1 see 3823 of 1957 (Upatov).

537.227 3435

Contribution to the Theory of Ferroelectric Polarization Curves—N. S. Akulov. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 1085–1087; April, 1959.) The principles of the polarization theory for single and many-domain ferroelectric crystals, with one or two Curie points, are developed for weak, strong, and medium fields. In contrast to magnetization-curve theory, rotation can be neglected in the first approximation.

537.311.33:537.29:538.63 3436

Electron Resonance in Crossed Electric and Magnetic Fields—I. M. Lifshits and M. I. Kaganov. (*Zh. eksp. teor. Fiz.*, vol. 37, pp. 555–556; August, 1959.) A note on the electric-field dependence of the period of revolution of an electron in a semiconductor in crossed fields.

537.32 3437

Thermoelectric Phenomena in Strong Magnetic Fields in Metals Possessing Various Fermi Surfaces—Yu. A. Bychkov, L. E. Gurevich, and G. M. Nedlin. (*Zh. eksp. teor. Fiz.*, vol. 37, pp. 534–539; August, 1959.) The asymptotic variation of the thermoelectric force, Peltier and Thomson coefficients for metals with closed Fermi surfaces and open surfaces of the "corrugated cylinder" and "space net" types is investigated on the basis of the quasi-classical theory of kinetic phenomena [1411 of 1957 (Lifshits, *et al.*)].

537.533 3438

Determination of the Shape of a Relativistic Electron Beam—N. L. Tsintsadze. (*Zh. Tekh. Fiz.*, vol. 29, pp. 24–26; January, 1959.) Brief mathematical analysis of an axially symmetric thin electron beam, the cross section of which varies along the axis of the beam, the movement of the electron beam being assumed adiabatic.

537.533 3439

Determination of Trajectories of Relativistic Charged Particles in Electric and Magnetic Fields by the Adams Method of Finite Differences—N. I. Shtepa. (*Zh. Tekh. Fiz.*, vol. 29, pp. 120–127; January, 1959.)

537.533:537.56 3440

Plasma Oscillations of a Large Number of Electron Beams—J. M. Dawson. (*Phys. Rev.*, vol. 118, pp. 381–389; April 15, 1960.) Longitudinal oscillations of a large number of electron beams are investigated. The problem of passing to the limit of a continuous velocity distribution is considered in detail.

537.56 3441

Electric Field Distributions in an Ionized Gas: Part 2—B. Mozer and M. Baranger. (*Phys. Rev.*, vol. 118, pp. 626–631; May 1, 1960.) The method described earlier [814 of March (Baranger and Mozer)] is used to calculate the probability distribution of the LF component of the field at a neutral point, and also, the distributions of the LF component at an ion and the HF component at an electron.

537.56 3442

Transport Phenomena in Slightly Ionized Gases: High Electric Fields—M. S. Sodha. (*Phys. Rev.*, vol. 118, pp. 378–381; April 15, 1960.) An investigation of the variation with electric field of a number of transport properties in the presence of a magnetic field perpendicular to the electric field and temperature gradient in the gas. For the case of low electric fields, see 1550 of May.

537.56 3443

The Equation of State of a Plasma—A. A. Vedenov and A. I. Larkin. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 1133–1142; April, 1959.) The free energy of a completely ionized gas is given in the form of an expansion in terms of the density of interacting particles.

537.56:537.533 3444

Anomalous Electron Scattering and the Excitation of Plasma Oscillations—M. D. Gabovich and L. L. Pasechnik. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 1025–1033; April, 1959.) An investigation of the interaction of an electron beam with an independently formed plasma. The observed data can be explained by assuming that the electrons form clusters which interact coherently with the plasma.

537.56:538.63 3445

Oscillations of a Plasma Beam in a Magnetic Field at Frequencies Close to the Cyclotron Frequency—V. N. Lazukin. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 969–975; April, 1959.) A method is described by which plasma oscillations have been observed in hydrogen at frequencies lower than the ion cyclotron frequencies. The absorption of RF power by ions which rotate in the magnetic field is detected by the change in Q of an oscillator circuit.

537.56:538.63 3446

Hydromagnetic Waves in Plasmas—W. O. Schumann. (*Z. angew. Phys.*, vol. 11, pp. 259–264; July, 1959.) A theoretical investigation of wave propagation along and across a strong magnetic field in a plasma, taking account of electron diffusion is made. See also 4179 of 1959, and back references.

538.12 3447

Note on Perturbation of a Uniform Magnetic Field by a Cube of Magnetic Material—F. Partovi. (*IBM J. Res. & Dev.*, vol. 4, pp. 357–358; July, 1960.

538.221 3448

On the Theory of Weak Ferromagnetism—E. A. Turov. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 1254–1258; April, 1959.)

538.221:537.122 3449

Ferromagnetism of an Electron Gas—M. Shimizu. (*J. Phys. Soc. Japan*, vol. 15, pp. 376–406; March, 1960.) A study is made on the basis of the Bohm-Pines collective description of electron interactions.

538.221:538.569.4 3450

Excitation of Spin Waves in Ferromagnets: Eddy-Current and Boundary-Condition Effects—P. Pincus. (*Phys. Rev.*, vol. 118, pp. 658–664; May 1, 1960.) The boundary condition for transverse magnetization is derived when there is a surface anisotropy field. Spin-wave modes can be excited by a uniform RF field, and the power absorbed in each mode in an insulator can be calculated. The eddy currents have a large effect only on long wavelength spin waves.

538.222:538.569.4 3451

Theory of Paramagnetic Resonance in Systems containing Two Kinds of Magnetic Moment—A. A. Kokin and G. V. Skrotskiĭ. (*Zh. eksp. teor. Fiz.*, vol. 37, pp. 482–489; August, 1959.) Equations are derived for the partial magnetizations in a weak variable magnetic field. Relaxation times and resonance-frequency shift are calculated.

538.222:538.569.4 3452

Spin-Lattice Relaxation Effects Observed in the Continuous Power Saturation of Paramagnetic Lines—G. V. Marr and P. Swarup. (*Canad. J. Phys.*, vol. 38, pp. 495–503; March, 1960.) The dependence of the conventional saturation parameter on the incident microwave power is considered for Lorentzian-shaped paramagnetic lines and applied to a study of the $(-\frac{1}{2}\rightarrow+\frac{1}{2})$ transitions of Cr^{+++} in $K_3Co(CN)_6$ and Gd^{+++} in $La(C_2H_5SO_4)_3\cdot 9H_2O$ at 9kMc and 4.2° K.

538.3 3453

Reciprocity Theorems for Electromagnetic Fields whose Time Dependence is Arbitrary—W. J. Welch. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 68–73; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

538.566 3454

Multilayer Absorbers for Electromagnetic Waves—H. J. Schmitt. (*Z. angew. Phys.*, vol. 11, pp. 335–339; September, 1959.) The absorbing power of structures which consist of two or more layers of loss-free dielectric and thin resistive material arranged alternately is discussed. The reflection coefficients are calculated for a number of such structures, which are treated as resonance absorbers [see, *e.g.*, 1846 of 1959 (Schmitt and Futtermenger)].

538.566:535.42 3455

High-Frequency Diffraction of Plane Waves by an Infinite Slit for Grazing Incidence—S. R. Seshadri and T. T. Wu. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 37–42; January, 1960. Abstract PROC. IRE, vol. 48, pt. 1, p. 1189; June, 1960.)

538.566:535.42 3456

High-Frequency Diffraction of Electromagnetic Waves by a Circular Aperture in an Infinite Plane Conducting Screen—S. R. Seshadri and T. T. Wu. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 27–36; January, 1960.) Abstract, PROC. IRE, vol. 48, pt. 1, p. 1189; June, 1960.)

538.566:537.56 3457

Reflection of an Electromagnetic Wave from a Plasma Moving through a Dielectric in a Constant Magnetic Field—Ya. B. Faĭnberg and V. S. Tkalich. (*Zh. Tekh. Fiz.*, vol. 29, pp. 491–496; April, 1959.) Mathematical treatment of the reflection of a retarded em wave from a moving plasma in a magnetic field directed parallel to the plasma motion. Expressions are derived for the frequency and amplitude of the reflected wave. For the Cherenkov

case [see 3035 of 1956 (Lampert)], the dependence of phase velocity on magnetic field intensity is determined, and two refracted waves are defined.

538.566:537.56 3458
Validity of the Theory of Double-Stream Amplification—D. T. Swift-Hook. (*Phys. Rev.*, vol. 118, pp. 1–5; April 1, 1960.) Recent misunderstandings concerning the validity of the original analysis of the interaction of moving ion or electron streams are clarified. See, *e.g.*, 3267 of 1959 (Piddington).

538.566.029.6:535.4 3459
Theory of Diffraction in Microwave Interferometry—D. M. Kerns and E. S. Dayhoff. (*J. Res. NBS*, vol. 64B, pp. 1–13; January–March, 1960.) The theory is applicable to Michelson and Fabry-Perot interferometers.

538.569 3460
Some Solutions for Electromagnetic Problems involving Spheroidal, Spherical, and Cylindrical Bodies—J. R. Wait. (*J. Res. NBS*, vol. 64B, pp. 15–32; January–March, 1960.) Results are given for the LF response to an oscillating magnetic dipole by conducting bodies of simple shape.

538.569.4:538.221 3461
Effect of Specimen Shape on Ferromagnetic Resonance in a Strong Radio-Frequency Field—G. V. Skrotskiĭ and Yu. I. Alimov. (*Zh. eksp. teor. Fiz.*, vol. 36, pp. 1267–1271; April, 1959.) An analysis of the exact solutions of the Landau-Lifshits equations is given. An expression is derived for the threshold field h_c above which instability in the motion of the magnetization vector begins. For resonance field strengths $h_o > h_c$, the height of the absorption peak decreases and its width increases.

538.569.4:621.3.018.41(083.74) 3462
Maser Oscillation on 6-6 Line of $N_{15}H_3$—I. Takahashi, T. Hashi, M. Yamano, M. Yamamoto, S. Suzuki, and T. Makita. (*J. Phys. Soc. Japan*, vol. 15, pp. 531–532; March, 1960.) The characteristics of maser oscillation on the $N^{15}H_3$ 6-6 line and the $N^{14}H_3$ 3-3 line are compared, and the mutual influence of $N^{15}H_3$ and $N^{15}H_3$ is studied. See also 2858 of August (Matsuura).

538.569.4:621.375.9 3463
Nuclear-Magnetic-Resonance Maser—H. Benoit. (*Ann. Phys.* (*Paris*), vol. 4, pp. 1439–1483; November/December, 1959.) A detailed investigation of a two-level weak-field maser-type self-oscillator based on nuclear-spin rotation in a circulating liquid under conditions of rapid adiabatic passage through resonance. Applications for magnetic-field and relaxation-time measurements are discussed. See 2521 of 1959 (Benoit, *et al.*).

538.569.4.029.65:538.222 3464
Broadband Millimeter-Wave Paramagnetic Resonance Spectrometer—J. B. Mock. (*Rev. Sci. Instr.*, vol. 31, pp. 551–555; May, 1960.) The specimen and waveguide which can be cooled down to 1.2°K by means of a cryostat are placed between close pole-pieces, giving high uniform magnetic fields. Design details and some measurements on zero-field splittings are given.

538.614 3465
New Analysis of the Polarization of Radiation and the Faraday Effect in Terms of Complex Vectors—K. C. Westford. (*J. Opt. Soc. Amer.*, vol. 49, pp. 717–723; July, 1959.) "The properties of the polarization ellipse are deduced in terms of the ratio of the Cartesian components of the complex electric vector of a beam of radiation by utilizing the Argand representation of a real two-dimensional vector as a complex number. The two components of a beam that are accepted and rejected by a polarizer or a radio antenna are considered as orthogonal components in the directions of two complex orthonormal vectors. The intensities of the corresponding components of a polarized beam are derived and represented on the Poincaré sphere. The methods are then applied to the important radio case of the Faraday effect in a uniform magneto-ionic medium. Finally, the measurable quantities characterizing a beam of partially polarized radiation are obtained from a diagonalization of the complex polarization tensor that specifies the beam."

539.124 3466
Absorption and Emission Spectra of an Electron in a One-Dimensional Deep Trap—B. S. Gourary and A. A. Maradudin. (*J. Phys. Chem. Solids*, vol. 13, pp. 88–104; May, 1960.)

539.2:537.311.33 3467
Generalization of Band Theory to Include Self-Energy Corrections—G. W. Pratt, Jr. (*Phys. Rev.*, vol. 118, pp. 462–467; April 15, 1960.) An investigation of the corrections to ordinary band theory caused by dielectric effects.

GEOPHYSICAL AND EXTRA-TERRESTRIAL PHENOMENA

523.164:621.396.677 3468
Maximum Angular Accuracy of Tracking a Radio Star by Lobe Comparison—R. Manasse. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 50–56; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

523.164:621.396.677 3469
The North-South Branch of the Great Interferometer of the Nancay Station—A. M. Malinge, E. J. Blum, M. Ginat, and M. Parise. (*Compt. rend. acad. sci., Paris*, vol. 249, pp. 2009–2011; November 16, 1959.) A brief description of the antenna system and receiving equipment which is being used in conjunction with the E-W interferometer [see 3644 of 1956 (Blum, *et al.*)] for a detailed study of the sun. The N-S branch operates at 169 Mc and has eight-parabolic mirrors 10m in diamcter spaced along a 770-m track. Provision is made for adjusting the declination of the beam.

523.164.3:523.4 3470
Microwave Absorption and Emission in the Atmosphere of Venus—A. H. Barrett. (*J. Geophys. Res.*, vol. 65, pp. 1835–1838; June, 1960.) Theoretical considerations suggesting that measurements in the mmλ range may give information on the atmosphere below the cloud layer of the planet are presented.

523.164.3:523.4 3471
The Source of Radiation from Jupiter at Decimetre Wavelengths: Part 2—Cyclotron Radiation by Trapped Electrons—G. B. Field. (*J. Geophys. Res.*, vol. 65, pp. 1661–1671; June, 1960.) Part 1: 468 of February.

523.164.3:523.4 3472
Ionospheric Modification of the Radio Emission from Jupiter—A. G. Smith, T. D. Carr, H. Bollhagen, N. Chatterton, and F. Six. (*Nature* (*London*), vol. 187, pp. 568–570; August 13, 1960.) The experiment of Gardner and Shain (3426 of 1958) has been repeated at 18 Mc using two identical receivers separated by 7040 km. Some correlation is found between records at the two sites but there is also evidence of serious distortion caused by terrestrial effects.

523.165 3473
Balloon Study of High-Altitude Radiations during the International Geophysical Year—J. R. Winckler. (*J. Geophys. Res.*, vol. 65, pp. 1331–1359; May, 1960.)

523.165 3474
Observations of the Van Allen Radiation Regions during August and September 1959: Part 1—R. L. Arnoldy, R. A. Hoffman, and J. R. Winckler. (*J. Geophys. Res.*, vol. 65, pp. 1361–1376; May, 1960.) Results are given of a preliminary analysis of data obtained from equipment flown on Explorer VI (1959 δ). In addition to the inner and outer zones, a stable intermediate zone was detected. The outer-zone intensity showed a large decrease following a sudden commencement but subsequently increased to be greatly in excess of its prestorm level. Comparison is made with other measurements of the radiation zones.

523.165 3475
Particle Fluxes in the Inner Radiation Belt—S. C. Freden and R. S. White. (*J. Geophys. Res.*, vol. 65, pp. 1377–1383; May, 1960.)

523.3:621.396.96 3476
Roughness of the Moon as a Radar Reflector—B. H. Briggs. (*Nature* (*London*), vol. 187, p. 490; August 6, 1960.) Radar observations of the distribution of slope of the lunar surface made by Hughes (3079 of September) are found to be in good agreement with optical measurements when the slope angle is greater than 1°.

523.3:621.396.96 3477
Radar Determination of the Scattering Properties of the Moon—F. B. Daniels. (*Nature* (*London*), vol. 187, p. 399; July 30, 1960.) Results obtained by Feinstein (2746 of 1954) are modified to take into account the curvature of an irregular reflecting surface, and applied to problems associated with scattering from a lunar surface which has two-dimensional corrugations. See also 3079 of September (Hughes).

523.53 3478
"Sporadic Shower" Properties of Very Small Meteors—P. B. Gallagher and V. R. Eshleman. (*J. Geophys. Res.*, vol. 65, pp. 1846–1847; June, 1960.) A long-term radar survey shows "sporadic" meteoric particles to be distributed in a vast number of relatively small groupings.

523.53:550.385.3 3479
Observed Magnetic Effects from Meteors—A. W. Jenkins, Jr., C. A. Phillips, Jr., and E. Maple. (*J. Geophys. Res.*, vol. 65, pp. 1617–1619; May, 1960) An analysis of subaudio fluctuations in the geomagnetic field indicates that meteoric activity and the average level of fluctuations are related.

523.53:621.396.96 3480
Experimental Studies of Meteor Echoes at 200 Mc—J. L. Heritage, S. Weisbrod, and W. J. Fay. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 57–61; January, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

523.755 3481
Properties of the Solar Wind during Sunspot Minimum—A. L. Licht. (*J. Geophys. Res.*, vol. 65, pp. 1397–1400; May, 1960.)

550.385.26 3482
Three-Dimensional Considerations for Current System of Geomagnetic Variations: Part 2—*Sq* Field—A. Nishida and N. Fukushima. (*Rept. Ionos. Space Res. Japan*, vol. 13, pp. 273–282; December, 1959.) Part 1: 3487 below.

550.385.26 3483
On the Position of the Focus of the Geomagnetic *Sq* Current System—M. Hasegawa. (*J. Geophys. Res.*, vol. 65, pp. 1437–1447; May,

1960.) A review of present knowledge of the *Sq* foci, discussion of the cause of changes in the *Sq* field, and a description of problems relating to the determination of the *Sq* focus are given. The foci in the potential fields of geomagnetic *Dq* variations may be taken as approximately indicating the positions of ionospheric *Sq* current foci.

550.385.37 3484
Concerning the Nature of Short-Period Magnetic Micropulsations—W. H. Campbell. (*J. Geophys. Res.*, vol. 65, pp. 1843–1845; June, 1960.) World-wide and local effects are discussed from the point of view that the pulsations originate within the ionosphere.

550.385.37:551.594.5:551.510.535 3485
Magnetic Micropulsations, Pulsating Aurora, and Ionospheric Absorption—W. H. Campbell. (*J. Geophys. Res.*, vol. 65, pp. 1833–1834; June, 1960.) Increases in cosmic-noise absorption at 27.6 Mc coincide with magnetic and auroral pulsations.

550.385.4 3486
Studies on Sudden Commencements of Geomagnetic Storms using IGY Data from United States Stations—S. Matsushita. (*J. Geophys. Res.*, vol. 65, pp. 1423–1435; May, 1960.) Report of a study of geomagnetic storm variations recorded over a period of one year from July 1, 1957, at a network of stations each separated by 360–510 km.

550.385.4(98) 3487
Three-Dimensional Consideration for Current System of Geomagnetic Variations: Part 1—Current Flow for Polar Elementary Storm within a Spherical Conducting Shell—N. Fukushima and R. Maeda. (*Rept. Ionos. Space Res. Japan*, vol. 13, pp. 267–272; December, 1959.) Continuation of 2348 of July (Fukushima).

551.507.362.1:523.165 3488
Preliminary Results from the Space Probe Pioneer V—C. Y. Fan, P. Meyer, and J. A. Simpson. (*J. Geophys. Res.*, vol. 65, pp. 1862–1863; June, 1960.) Cosmic-ray measurements show that, (*a*) Forbush decreases are not dependent on the presence of the earth's field as they occur out in space, (*b*) solar flare particles producing polar-cap absorption are not stored in the geomagnetic field, and (*c*) solar production of energetic electrons is not restricted to periods of flare activity.

551.507.362.1:550.37 3489
Characteristics of the Extraterrestrial Current System: Explorer VI and Pioneer V—E. J. Smith, P. J. Coleman, D. L. Judge, and C. P. Sonett. (*J. Geophys. Res.*, vol. 65, pp. 1858–1861; June, 1960.) Magnetometer measurements of the earth's field are consistent with a toroidal ring-current flowing in the magnetic equatorial plane at a distance of 9- to 10-earth radii. The internal radius of the toroid is 3-earth radii, and it carries a current of 5×10^6A.

551.507.362.1:550.380.8 3490
Some Preliminary Results of the Pioneer V Magnetometer Experiment—P. J. Coleman, Jr., C. P. Sonett, D. L. Judge, and E. J. Smith. (*J. Geophys. Res.*, vol. 65, pp. 1856–1857; June, 1960.) The sudden increase in magnetic field strength during solar activity suggests that collisionless magnetoacoustic waves may be formed at the boundary between the earth's field and the interplanetary field which is put at 13-earth radii.

551.507.362.2 3491
Induction Drag on a Large Negatively Charged Satellite Moving in a Magnetic-Field-Free Ionosphere—P. J. Wyatt. (*J. Geophys. Res.*, vol. 65, pp. 1673–1678; June, 1960.)

551.507.362.2 3492
The Influence of the Solar Radiation Pressure on the Motion of an Artificial Satellite—P. Musen. (*J. Geophys. Res.*, vol. 65, pp. 1391–1396; May, 1960.) Results of theoretical investigations of the radiation-pressure perturbation have been applied to orbital calculations for Vanguard 1 (1958β), producing significant perturbations in its perigee height. In the case of a satellite with a large area/mass ratio, the major terms introduced by solar radiation pressure can reduce the perigee distance at the rate of 1–2 km/day.

551.507.362.2 3493
Final Months and Descent of Satellite Sputnik 3—B. R. May and D. E. Smith. (*Nature (London)*, vol. 187, pp. 456–458; August 16, 1960.) A report of the progress of 1958 δ_2 from May, 1959, until its descent on April 6, 1960, and a brief discussion of the main features of its 692-day lifetime. For earlier progress see 141 of January.

551.507.362.2:551.510.535 3494
The Upper Earth Atmosphere according to Satellite Measurements—H. K. Paetzold. (*Naturwissenschaften*, vol. 46, pp. 416–422; July, 1959.) A review of results obtained from measurements on the first five earth satellites is given. About 50 references.

551.507.362.2:621.391.812.63 3495
First Results of Satellite Observations for Information on Wave Propagation—B. Beckmann. (*Nachrichtentech. Z.*, vol. 12, pp. 335–343; July, 1959.) A review of earlier radio and radar observations of earth satellites, and a summary of results obtained relating to ionospheric phenomena are given.

551.507.362.2:621.391.812.63:539.16 3496
Frequency Scintillations of Satellite Signals before and after the Argus Experiments—P. R. Arendt. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 73–77; January, 1960). Abstract, PROC. IRE, vol. 48, pt. 1, p. 1190; June, 1960.)

551.510.535 3497
A Convenient Method of getting Representative Ionospheric Heights—G. A. M. King. (*J. Geophys. Res.*, vol. 65, pp. 1623–1624; May, 1960.) A simple method for obtaining electron-density profiles is described. Monthly median *h'f* curves are drawn for each hour and then reduced to true height profiles by any standard method.

551.510.535 3498
The Joint Use of the Ordinary and Extraordinary Virtual-Height Curves in Determining Ionospheric Layer Profiles—L. R. O. Storey. (*J. Res. NBS*, vol. 64D, pp. 111–124; March/April, 1960.) An extension of Budden's matrix method for determining ionospheric layer profiles. Errors are reduced by using the extraordinary virtual-height curve, as well as the ordinary curve.

551.510.535 3499
The Formation of the D Region of the Ionosphere—N. Nicolet and A. C. Aikin. (*J. Geophys. Res.*, vol. 65, pp. 1469–1483; May, 1960.) A review of theoretical studies of the ionization process in the *D* region. It is possible to explain normal conditions of ionization by cosmic rays and Lyman α radiation. Conditions caused by solar flares must be explained by X rays. Above 85 km, the ionization is related to the formation of the *E* layer.

551.510.535 3500
Field-Aligned Ionization—J. Mawdsley and W. H. Ward. (*J. Geophys. Res.*, vol. 65, pp. 1828–1830; June, 1960.) Field-aligned ionization in the *E* and *F* regions may result from leakage of particles from the radiation belts.

551.510.535 3501
Longitude Effect in Temperate-Zone Sporadic E and the Earth's Magnetic Field—L. H. Heisler and J. D. Whitehead. (*Nature (London)*, vol. 187, pp. 676–677; August 20, 1960.) A correlation in both northern and southern temperate zones between the magnitude of the horizontal component of the earth's magnetic field and the percentage of time that f_0-E_s is >5 Mc, is shown to exist. This is considered with reference to the longitude effect investigated by Smith and Finney (2736 of August), and it is suggested that the movement of ionization by winds or electric fields is the most likely source of E_s.

551.510.535 3502
Drift Measurements at Kjeller on the Ionospheric F Region—J. Becken and B. Maehlum. (*J. Geophys. Res.*, vol. 65, pp. 1485–1488; May, 1960.) Results are given of measurements made at Kjeller, Norway, mostly during January and February, 1959. The mean drift velocity was found to be 140 msec towards the west at night and the east during the day. Some measurements on sporadic *F* echoes are reported.

551.510.535 3503
Inferred Temperature Variations at the F_2 Peak—S. J. Bauer. (*J. Geophys. Res.*, vol. 65, pp. 1685–1690; June, 1960.) The scale height at the F_2 peak has been estimated, using data on the total electron content of the ionosphere obtained from the Faraday rotation of lunar radio echoes and data obtained from vertical-incidence soundings. The scale height varies with time, probably as a result of temperature variations. The observed scale heights are in good agreement with a model atmosphere based on rocket and satellite observations.

551.510.535:539.16 3504
Vertically Travelling Shock Waves in the Ionosphere—F. B. Daniels, S. J. Bauer, and A. K. Harris. (*J. Geophys. Res.*, vol. 65, pp. 1848–1850; June, 1960.) Ionospheric phenomena at vertical incidence observed in the vicinity of ground-based atomic explosions are interpreted in terms of acoustic and hydromagnetic wave propagation.

551.510.535:551.593.9 3505
Oxygen Red Line in the Night Airglow and Ionospheric F_2 Region—M. Huruhata, T. Nakamura, H. Tanabe, and T. Tohmatsu. (*Rep. Ion. Space Res. Japan*, vol. 13, pp. 283–289; December, 1959.) A statistical analysis of spectroscopic data indicates that the intensity of the red radiation and the maximum electron density of the F_2 region are related.

551.510.535:551.594.5 3506
Some Characteristics of the D-Region Ionization during Auroral Activity—W. Stoffregen, H. Derblom, and A. Omholt. (*J. Geophys. Res.*, vol. 65, pp. 1699–1704; June, 1960.) In north Sweden, *D*-layer ionization during aurorae is closely associated with sudden increases in auroral activity and extends down to about 65 km. The latitude variations of this layer are discussed, and it seems likely that the ionization is caused by primary electrons and secondary X rays.

551.510.535:621.391.812.63.029.45 3507
V. L. F. Propagation Effects of a D-Region Layer Produced by Cosmic Rays—Moler. (See 3638.)

551.510.536 3508
A New Mechanism for Accelerating Electrons in the Outer Ionosphere—R. A. Helliwell and T. F. Bell. (*J. Geophys. Res.*, vol. 65,

pp. 1839–1842; June, 1960.) An experiment is outlined to produce high-energy electrons in the ionosphere using em radiation from a ground-based LF transmitter. The principle is adaptable for use in a satellite or space probe.

551.510.536:621.391.812 3509
Scattering of Radio Waves by Electrons above the Ionosphere—E. E. Salpeter. (*J. Geophys. Res.*, vol. 65, pp. 1851–1852; June, 1960.) By measuring the over-all intensity of the scattered spectrum it should be possible to estimate electron density and temperature; if better resolution were possible, the atomic weights of the ions present could be obtained.

551.594.2 3510
Lightning and Charge Storage—E. J. Workman, M. Brook, and N. Kitagawa. (*J. Geophys. Res.*, vol. 65, pp. 1513–1517; May, 1960.) A lightning flash consisting of 54 current surges of which 26 were leader/return-stroke combinations and having a total duration of two seconds is described.

551.594.221:551.508.86 3511
The Design and Use of Instruments for Counting Local Lightning Flashes—F. Horner. (*Proc. IEE*, vol. 107, pt. B, pp. 321–330; July, 1960.) The instrument adopted as a standard for assessing local activity triggers on a field change of 3 vpm. The need for an international standard is stressed.

551.594.5 3512
Theory of Auroral Structures—E. R. Harrison. (*Nature* (*London*), vol. 187, pp. 383–386; July 30, 1960.) A simplified theoretical examination of the structure of visual aurorae.

551.594.5 3513
Movement of Auroral Ionization—J. Mawdsley. (*J. Geophys. Res.*, vol. 65, pp. 1825-1827; June, 1960.) A note on the apparent discrepancies between movements deduced from visual and radar observations.

551.594.5 3514
Observations of the Aurora Australis at New Zealand Antarctic Stations during I.G.Y.—T. Hatherton and G. G. Midwinter. (*J. Geophys. Res.*, vol. 65, pp. 1401–1411; May, 1960.)

551.594.5:550.385 3515
Correlation Studies of Radio-Aurora, Magnetic, and Earth-Current Disturbances—B. K. Bhattacharyya. (*Canad. J. Phys.*, vol. 38, pp. 624–637; May, 1960.) The diurnal variation of the rate of auroral echo occurrence appears to be similar to that of moderate perturbations of H. A relation is suggested between magnetic perturbations and the reversal of direction of auroral ionization drift from west to east at about local midnight.

551.594.5:621.396.96 3516
On the Geometry of Radio Reflections from Aurora—P. A. Forsyth. (*Canad. J. Phys.*, vol. 38, pp. 593–603; May, 1960.) The observed characteristics of auroral radar echoes are explained by assuming that echoes arise by means of volume scattering from configurations of ionization similar to those observed visually. The importance of allowing for the radar pulse length and antenna beam width is stressed.

551.594.5:621.396.96 3517
The Association of Visible Auroral Forms with Radar Echoes—G. F. Lyon. (*Canad. J. Phys.*, vol. 38, pp. 385–389; March, 1960.) An extension of earlier work [2255 of 1959 (Lyon and Kavadas)] to detect a more detailed correlation between echoes and specific visible auroral forms.

551.594.5:621.396.96 3518
An Analysis of some Statistical Properties of Auroral Radar Reflections and their Relationships to the Detection Capabilities of the Radar—A. G. McNamara. (*Canad. J. Phys.*, vol. 38, pp. 425–438; March, 1960.) Expressions are derived from an analysis of protracted observations at 48.5 Mc, relating radar parameters to echo statistics.

551.594.6 3519
Very-Low-Frequency Atmospherics—H. König. (*Z. angew. Phys.*, vol. 11, pp. 264–274; July, 1959.) Report on measurements of atmospherics in the frequency range 0.5–25 cps by means of specially designed equipment. The results are discussed with reference to measurements made by others. A sinusoidal signal of about 9 cps was observed which may be caused by resonance of the system earth-ionosphere.

551.594.6 3520
Geomagnetic Noise at 230 kc/s—R. L. Dowden. (*Nature* (*London*), vol. 187, pp. 677–678; August 20, 1960.) Results of observations made at 4, 9, 70 and 230 kc are compared with estimates of the noise intensity to be expected from a Cherenkov radiation mechanism.

551.594.6:550.385.4 3521
Geomagnetic Disturbances and 5 Kilocycles-per-Second Electromagnetic Radiation—G. R. A. Ellis. (*J. Geophys. Res.*, vol. 65, pp. 1705–1710; June, 1960.) Out of 97 isolated noise bursts at 5 kc in Australia, 43 were associated with positive bays in the H component of the earth's field. Noise bursts followed the occurrence of 30 magnetic sudden commencements with a delay of 8.5 hours, but on four occasions there was no delay. See 2384 of July.

LOCATION AND AIDS TO NAVIGATION

621.396.93+621.396.96 3522
Future Trends of Radio and Radar Navigation—C. Williams. (*J. Brit. IRE*, vol. 20, pp. 417–428; June, 1960.) The requirements of marine and air navigation systems are compared, and present systems are discussed. New techniques including inertial navigation, a global VLF system, and the use of radio sextants and earth satellites are mentioned. Proposals for introducing automation in air traffic control are noted.

621.396.962.25 3523
Analysis of a Frequency-Modulated Continuous-Wave Ranging System—A. J. Hymans and J. Lait. (*Proc. IEE*, vol. 107, pt. B, pp. 365–372; July, 1960.) A re-examination of the problem for moving targets leads to proposals for a system overcoming the restrictions of the single-line beat-note spectrum assumed by Gnanalingam (3219 of 1954).

621.396.962.38 3524
Solid-State Modulator Feeds Subminiature Transponder—L.. Diven. (*Electronics*, vol. 33, pp. 48–51; July 1, 1960.) High delay stability provides a range accuracy within about one yard for distances up to 200 miles.

621.396.969.3 3525
Radar Target Classification by Polarization Properties—J. R. Copeland. (PROC. IRE, vol. 48, pp. 1290-1296; July, 1960.) The polarization may be studied by using a rotating linear antenna, and results enable the target to be classified. The polarization properties of model targets are derived.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 3526
Photoconductivity in Ternary V-VI-VII Compounds—R. Nitsche and W. J. Merz (*J. Phys. Chem. Solids*, vol. 13, pp. 154–155; May, 1960.) An experimental study of some photoelectric properties of single crystals.

535.215:546.48'221 3527
Photoemission in the Photovoltaic Effect in Cadmium Sulphide Crystals—R. Williams and R. H. Bube. (*J. Appl. Phys.*, vol. 31, pp. 968-978; June, 1960.) The photovoltaic current in Cu-CdS and related systems, with undiffused metal/semiconductor junctions, is shown to result from photoemission of electrons from the Cu metal into the CdS crystal. The conditions required for this process, and factors affecting the efficiency of such cells, are discussed.

535.215:546.48'221 3528
Etch Pits and Dislocations in Cadmium Sulphide Crystals—J. Woods. (*Brit. J. Appl. Phys.* vol. 11, pp. 296–302; July, 1960.)

535.215:621.382.2 3529
Photoeffects in Nonuniformly Irradiated p-n Junctions—G. Lucovsky. (*J. Appl. Phys.*, vol. 31, pp. 1088–1095; June, 1960.) A theory for interpreting the photoeffects is derived by application of the usual continuity and diffusion equations. Confirmatory experimental results are given.

535.215-15 3530
Cooled Photoconductive Infrared Detectors—W. Beyen, P. Bratt, H. Davis, L. Johnson, H. Levinstein, and A. Macrae. (*J. Opt. Soc. Amer.*, vol. 49, pp. 686–692; July, 1959.) PbTe and two types of Au-doped Ge infrared detectors have been constructed, and their characteristics analysed.

535.37:546.47'221 3531
On the Infrared Emission in ZnS:Cu—Effect of Sulphur Pressure and Aluminum—E. F. Apple and J. S. Prener. (*J. Phys. Chem. Solids*, vol. 13, pp. 81–87; May, 1960.)

535.37:546.47'221 3532
The Structure of Luminescence Centres in Zinc Sulphide Phosphors containing Activators—N. Riehl and H. Ortmann. (*Ann. Phys.* (*Lpz.*), vol. 4, pp. 3–14; July 21, 1959.) The reversibility of blue and green Cu centers at low temperatures is investigated experimentally.

535.37:546.48'221 3533
Temperature Dependence of Edge Emission in Cadmium Sulphide—D. C. Reynolds. (*Phys. Rev.*, vol. 118, pp. 478–479; April 15, 1960.)

535.376 3534
Variation of the Electroenhancement Effect with the Frequency of the Electric Field—J. Mattler. (*Compt. rend. acad. sci., Paris*, vol. 249, pp. 2051–2053; November 16, 1959.) When the frequency of the applied field is increased from 20 to 4100 cps, the electroenhancement effect decreases rapidly, whatever method of phosphor excitation is used.

535.376:546.47'221 3535
Influence of Temperature on the Spectral Composition of the Zinc Sulphide Luminescence—I. Soudek. (*Brit. J. Appl. Phys.*, vol. 11, pp. 289–292; July, 1960.) Experimental results clearly show the temperature-dependent changes of the emission band, and an explanation is given.

537.226:621.317.335.3.029.65 3536
Determination of the Dielectric Properties of Low-Loss Ceramics at Q-Band Frequencies—J. M. Free and G. B. Walker. (*Proc. IEE*, vol. 107, pt. B, pp. 354–356; July, 1960.) The H_{10} made in a cylindrical cavity leads to low wall losses ($Q \sim 15000$). Loss tangents of disk specimens of Mg titanate and TiO_2 ceramics were about 0.0003, with relative permittivities of 14 and 80, respectively.

537.226:621.319.2 3537
Investigations on Electrets of Carnauba Wax and the Origin of their Heterocharge—J. van Calker and W. Frölich. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 216–224; July 21, 1959.) The results of experimental investigations of heterocharge and its relaxation time are interpreted on the basis of MacDonald's theory of ac space-charge polarization effects (1024 of 1954). For investigations of homocharge see 2405 of July (van Calker and van der Linde).

537.227 3538
Ferroelectric Behaviour of Barium Zirconium Metaniobate—G. Goodman. (*J. Amer. Ceram. Soc.*, vol. 43, pp. 105–113; February, 1960.) Physical properties consistent with the idea of a ferroelectric-paraelectric transition of higher than second order have been observed in $BaNb_{1.5}Zr_{0.25}O_{5.25}$.

537.227 3539
The Relation between Structure and Ferroelectricity in Lead Barium and Barium Strontium Niobates—M. H. Francombe. (*Acta Cryst.*, vol. 13, pt. 2, pp. 131–140; February 10, 1960.)

537.227 3540
Bias in Ferroelectric Colemanite—E. Fatuzzo. (*J. Appl. Phys.*, vol. 31, pp. 1029–1034; June, 1960.) An intrinsic bias exists above the Curie point and is responsible for the pyroeffect observed there. See also 2964 of 1954 (Devonshire).

537.227 3541
Dielectric Behaviour of Single Crystals of Triglycine Sulphate (TGS) from 1 kc to 2500 Mc—A Lurio and E. Stern. (*J. Appl. Phys.*, vol. 31, pp. 1125–1126; June, 1960.) Above the Curie point (~49°C) the dielectric constant follows a Curie-Weiss law, the constant for which is determined at various frequencies in the range.

537.227:546.431'824–31 3542
Space Charge Fields in $BaTiO_3$—S. Triebwasser. (*Phys. Rev.*, vol. 118, pp. 100–105; April 1, 1960.) Observations have been made, in the presence of dc electric field bias, of hysteresis loops, and above the Curie point, of double loops, the Kerr electro-optical effect and the capacitance. The results are used to study the effect of space-charge surface layers on the bulk field.

537.228.1 3543
An Alternative Transformation for the Elastic and Piezoelectric Constants of Anisotropic Media—R. Bechmann. (*Acta Cryst.*, vol. 13, pt. 2, pp. 110–113; February 10, 1960.) A useful form of transformation has been derived based on a more general definition of the Cristoffel moduli.

537.228.1:621.372.412 3544
Excitation of Piezoelectric Plates by Use of a Parallel Field with Particular Reference to Thickness Modes of Quartz—R. Bechmann. (Proc. IRE, vol. 48, pp. 1278–1280; July, 1960.) Quartz AT-cut crystals excited by a parallel field are of practical interest for high-precision frequency control. Data on the equivalent electric circuit of a 1-Mc oscillator, excited by a parallel field, are given.

537.228.2 3545
Jump Phenomena in Resonance Curve of Ferroelectric Ceramics—K. Negishi. (*J. Phys. Soc. Japan*, vol. 15, p. 534; March, 1960.) Resonance curves obtained when measuring the amplitude dependence of the mechanical Q are described. The jump or hysteresis phenomenon occurred only in high-Q specimens.

537.311.33 3546
The Theory of Surface Currents on Semiconductor Surfaces—E. Groschwitz and R. Ebhardt. (*Z. angew. Phys.*, vol. 11, pp. 296–304; August, 1959.) A model of the surface currents is derived on the basis of the results of inversion-layer investigations (1653 of May). The characteristics of surface current are calculated for a point contact diode and an alloy-junction transistor, and the effects of an externally applied potential are considered.

537.311.33:537.311.62 3547
Skin Effect in Semiconductors—A. H. Frei and M. J. O. Strutt. (Proc. IRE, vol. 48, pp. 1272–1277; July, 1960.) Formulas for flat plates are derived for field distribution, impedance, and power losses. Good agreement with experiment was found for the impedance of Ge at microwave frequencies.

537.311.33:537.32 3548
Thermo-EMF and Thermal Conduction of Group III-V Compounds and their Mixed Crystals—H. Weiss. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 121–131; July 21, 1959.) Measurements of differential thermo-EMF as a function of $1/T$ were made with the apparatus described earlier (1818 of 1957) on n- and p-type specimens of $In(As_yP_{1-y})$. The electron mobility and thermal conduction were also investigated to determine the relation between the scattering mechanism of electrons and that of phonons. A comparison is made with results obtained on other Group III-V compounds.

537.311.33:546.24 3549
Mechano-electrical Properties of Tellurium Single Crystals—G. Syrbe. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 132–139; July 21, 1959.) Investigation of the influence of expansion in the direction of the hexagonal axis on the electrical conductivity and thermo EMF parallel to the axis in needle-shaped crystals.

537.311.33:546.28+546.289 3550
Exciton Energy Levels in Germanium and Silicon—T. P. McLean and R. Loudon (*J. Phys. Chem. Solids*, vol. 13, pp. 1–9; May, 1960) "The energies of the lowest-lying levels of the direct excitron in Ge and the indirect excitons in Ge and Si have been calculated in the effective-mass approximation by a variational procedure. These energies are compared with the experimental values obtained from optical absorption experiments."

537.311.33:[546.28+546.289 3551
Internal Field Emission at Narrow Silicon and Germanium p-n Junctions—A. G. Chynoweth, W. L. Feldmann, C. A. Lee, R. A. Logan, G. L. Pearson, and P. Aigrain. (*Phys. Rev.*, vol. 118, pp. 425–434; April 15, 1960.) A detailed study has been made of the reverse characteristics of several alloyed p-n junctions with breakdown voltages in the range of about 0.1–0.8 V to examine the implications of the theoretical form of the tunneling probability, particularly as modified to take into account phonon cooperation.

537.311.33:546.28 3552
The Adsorption of Oxygen on Clear Silicon Surfaces—M. Green and K. H. Maxwell. (*J. Phys. Chem. Solids*, vol. 13, pp. 145–150; May, 1960.) Experimental data for the pressure range 30–200 μ Hg are discussed.

537.311.33:546.28 3553
Preparation of Hyper-pure Silicon for Semiconductor Use—H. Baba and H. Araki. (*Rep. Elect. Commun. Lab., Japan*, vol. 7, pp. 408–413; November, 1959.) A brief description of a pilot-plant system for the production of pure Si by the iodide process is given.

537.311.33:546.28 3554
Epitaxial Growth of Silicon—E. S. Wajda, B. W. Kippenhan, and W. H. White. (*IBM J. Res. & Dev.*, vol. 4, pp 288–295; July, 1960.) A study of the growth of Si layers on a Si substrate at 800°C by the dissociation of SiI_2.

537.311.33:546.28 3555
Impurity Introduction during Epitaxial Growth of Silicon—R. Glang and B. W. Kippenhan. (*IBM J. Res. & Dev.*, vol. 4, pp. 299–301; July, 1960.) Experiments show that certain Group III and V elements such as B, P, As and Sb can be transported and co-deposited by the SiI_2 process (see 3554 above).

537.311.33:546.28 3556
Liquid-Solid Interface Shape observed in Silicon Crystals Grown by the Czochralski Method—W. D. Edwards. (*Canad. J. Phys.*, vol. 38, pp. 439–443; March, 1960.)

537.311.33:546.28 3557
***p* Layers on Vacuum-Heated Silicon**—F. G. Allen, T. M. Buck, and J. T. Law. (*J. Appl., Phys.*, vol. 31, pp. 979–985, June, 1960.) Evidence is given that the p-type layer is formed by boron which comes from the borosilicate glass envelope. See also 559 of February (Allen, *et al.*).

537.311.33:546.28 3558
Low-Temperature Impurity Conduction in *n*-Type Silicon—K. R. Atkins, R. Donovan, and R. H. Walmsley. (*Phys. Rev.*, vol. 118, pp. 411–414, April 15, 1960.) Hall coefficients and conductivities were measured for specimens with impurity concentration of P about $10^{17}/cm^3$ and of B about $10^{15}/cm^3$. Results which are extremely sensitive to the density of minority impurities are discussed in relation to theory.

537.311.33:546.28 3559
High-Frequency Measurement of Carrier Lifetime in Silicon Single Crystals—W. Keller. (*Z. angew. Phys.*, vol. 11, pp. 351–352; September, 1959.) The usual photoelectric method is modified so as to permit the contactless testing of the Si rod. A 35-Mc alternating voltage is applied capacitively to the specimen under test. Results obtained on four different rods agree well with those of dc tests.

537.311.33:546.28:538.63 3560
Galvamagnetic Effects in *n*-Type Silicon—W. E. Krag. (*Phys. Rev.*, vol. 118, pp. 435–450; April 15, 1960.) Detailed comparisons of experimental and theoretical data are given.

537.311.33:[546.289+546.681'19 3561
Germanium-Gallium Arsenide Heterojunctions—R. L. Anderson. (*IBM J. Res. & Dev.*, vol. 4, pp. 283–287; July, 1960.) Abrupt monocrystalline junctions between two different semiconductor materials (heterojunctions) have been made by depositing Ge epitaxially on GaAs substrates; electrical characteristics of these junctions are summarized.

537.311.33:546.289 3562
Vapour-Deposited Single-Crystal Germanium—R. P. Ruth, J. C. Marinace, and W. C. Dunlap, Jr. (*J. Appl. Phys.*, vol. 31, pp. 995–1006; June, 1960.) Details are given of the preparation and properties of Ge layers formed on single-crystal Ge substrates by thermal decomposition of GeI_2.

537.311.33:546.289 3563
Epitaxial Vapour Growth of Ge Single Crystals in a Closed-Cycle Process—J. C. Marinace (*IBM J. Res. & Dev.*, vol. 4, pp. 248–255; July, 1960.) Detailed description of a GeI_2 reaction process for the deposition of pure Ge epitaxially upon Ge seeds in a single closed

tube at a temperature of 400°C. Modifications to the process to allow the growing deposits to be doped with suitable impurities are also discussed.

537.311.33:546.289 3564
Dislocation Content in Epitaxially Vapour-Grown Ge Crystals—H. S. Ingham, Jr., and P. J. McDade. (*IBM J. Res. & Dev.*, vol. 4, pp. 302–304; July, 1960.)

537.311.33:546.289 3565
Radio-Tracer Studies of the Incorporation of Iodine into Vapour-Grown Ge—W. E. Baker and D. M. J. Compton. (*IBM J. Res. & Dev.*, vol. 4, pp. 269–274; July, 1960.) Results indicate that the large amount of iodine present during the vapour-growth process has no significant effect on the properties of the Ge device produced.

537.311.33:546.289 3566
Incorporation of As into Vapour-Grown Ge—W. E. Baker and D. M. J. Compton. (*IBM J. Res. & Dev.*, vol. 4, pp. 275–279; July, 1960.) Radioactive-tracer methods showed that all the As incorporated into the Ge was electrically active, at least for material grown on a (211) seed.

537.311.33:546.289 3567
Incorporation of Au into Vapour-Grown Ge—W. E. Baker and D. M. J. Compton. (*IBM J. Res. H Dev.*, vol. 4, pp. 296–298; July, 1960.) Results of experiments using radioactive Au wire as a doping agent indicate that not all the Au incorporated into Ge using the closed-tube method is electrically active.

537.311.33:546.289 3568
Plastically Deformed Germaniun *p-n* Junctions—M. Bernard and B. Leduc. (*J. Phys. Chem. Solids*, vol. 13, pp. 168–169; May, 1960. In French.) It is shown that the phenomena of soft breakdown may be attributed to dislocations.

537.311.33:546.289:538.6 3569
Theory of the Valence Band Structure of Germanium in an External Magnetic Field—R. F. Wallis and H. J. Bowlden. (*Phys. Rev.*, vol. 118, pp. 456–461; April 15, 1960.) The quantum effects investigated may lead to observable phenomena in magneto-optic studies of Ge and similar semiconductors.

537.311.33:546.289:539.234 3570
Temperature Dependence of the Conductivity of Amorphous Vapour-Deposited Germanium Films—H. Richter and R. Schneider. (*Z. angew. Phys.*, vol. 11, pp. 277–283; July, 1959.) Conductivity measurements on Ge films prepared under different conditions were made in high vacuum for the temperature range 20°–600° C. The transition from amorphous to crystalline state is particularly studied.

537.311.33:546.4'241 3571
Solid Solution in $A^{II}B^{VI}$ Tellurides—J. C. Woolley and B. Ray. (*J. Phys. Chem. Solids*, vol. 13, pp. 151–153; May, 1960.) A determination of the lattice parameter variations in the three systems CdTe-HgTe, CdTe-ZnTe and HgTe-ZnTe.

537.311.33:546.681'18 3572
Infrared Lattice Absorption of GaP—D. A. Kleinman and W. G. Spitzer. (*Phys. Rev.*, vol. 118, pp 110–117; April 1, 1960.) Transmission and reflectivity measurements on GaP in the wavelength region 1–40 μ have been analysed.

537.311.33:546.681'18 3573
Anharmonic Forces in the GaP Crystal—D. A. Kleinman. (*Phys. Rev.*, vol. 118, pp. 118–127; April 1, 1960.) The infrared properties and the thermal expansion of GaP are treated on the basis of a simple model of the anharmonic forces. See also **3572** above.

537.311.33:546.681'19 3574
Properties of *p*-Type GaAs prepared by Copper Diffusion—F. D. Rosi, D. Meyerhofer, and R. V. Jensen. (*J. Appl. Phys.*, vol. 31, pp. 1105–1108; June, 1960.)

537.311.33:546.681'19 3575
Gallium Arsenide as a Semi-insulator—J. W. Allen. (*Nature (London)*, vol. 187, pp. 403–405; July 30, 1960.) Semi-insulators such as GaAs, in which a shallow donor is autocompensated by a deep acceptor, should exhibit pronounced hole trapping effects but little electron trapping.

537.311.33:546.682 3576
Semiconducting Indium Compounds—(*Electronic Engrg.*, vol. 32, pp. 373–375; June, 1960.) The preparation, properties and applications are briefly reviewed, and attention is drawn to the important range of ternary compounds.

537.311.33:546.682'86 3577
Properties of *p*-Type InSb in Pulsed High Electric Fields—M. C. Steele and M. Glicksman. (*Phys. Rev.*, vol. 118, pp. 474–477; April 15, 1960.) Experimental results at 77°K are described. Electron-hole pair creation occurs at electric fields >700 v/cm. The nature of the carrier initiating the pair creation is examined in detail.

537.311.33:546.682'86:539.23 3578
Production and Properties of Thin Layers of Indium Antimonide—G. Bate and K. N. R. Taylor. (*J. Appl. Phys.*, vol. 31, pp. 991–994; June, 1960.) Layers of polycrystalline InSb 10-μ thick have been produced by suddenly squashing a molten drop of the substance between two optical flats. The properties of the specimens are described.

537.311.33:546.873'241 3579
Anisotropic Diffusion of Copper into Bismuth Telluride—R. O. Carlson. (*J. Phys. Chem. Solids*, vol. 13, pp. 65–70; May, 1960.)

537.311.33:621.382.23 3580
Surface Problems in Barrier-Layer Semiconductors—H. H. Plagemann. (*Nachrtech.*, vol. 9, pp. 292–295; July, 1959.) Changes arising from the adsorption of gases by the semiconductor surface and the effect of moisture of the surrounding atmosphere on *p-n*-junction characteristics are discussed. The resulting manufacturing difficulties are overcome by manipulating the etched material in a climatically controlled dry chamber until the semiconductor device is finally sealed.

537.311.33:621.382.23 3581
Surface Studies on *p-n* Junctions—I. Stelzer. (*J. Electronics Control*, vol. 8, pp. 39–57; January, 1960.) A theoretical and experimental study is made of the effect on the electrical characteristics of a semiconductor material of adsorbed chemicals such as sulphur dioxide and diethylamine.

537.311.33:621.391.822 3582
Intervalley Noise—P. J. Price. (*J. Appl. Phys.*, vol. 31, pp. 949–953; June, 1960.) "A theory is developed for the spectrum of electrical noise due to electron transitions between several quasi-isolated groups of states, in the general case where each group may carry part of an electric current. It is applied to the noise due to transitions between valleys of the conduction band of germanium, and the possibility of observing this noise is discussed using the data of Weinreich, Sanders, and White on the frequency of intervalley transitions [3373 of 1959]."

537.312.62:538.63 3583
Surface Impedance of a Superconductor in a Magnetic Field—G. Dresselhaus and M. S. Dresselhaus. (*Phys. Rev.*, vol. 118, pp. 77–91; April 1, 1960.) The surface impedance is calculated. Comparison between theory and experimental data yields information on the energy band structure.

537.312.62:538.63 3584
Speed of the Superconducting-Normal Transition in Tin Films—D. J. Oliver, M. J. Rayner, and E. H. Rhoderick. (*Nature (London)*, vol. 187, p. 492; August 6, 1960.) Experiments on films 250-μ wide 1000-Å thick, and 1-cm long showed the relaxation time for the destruction of superconductivity by a magnetic field above the critical value to be $<10^{-8}$ seconds.

537.312.62:621.318.57 3585
Thermal Propagation of a Normal Region in a Thin Superconducting Film and its Application to a New Type of Bistable Element—R. F. Broom and E. H. Rhoderick. (*Brit. J. Apply. Phys.*, vol. 11, pp. 292–296; July,1960.) The movement of the interphase boundary resulting from current heating in a partially superconducting film is analysed. For a critical value of current, no movement takes place. This is confirmed experimentally and applied to the design of a bistable element. See also *Phys. Rev.*, vol. 116, pp. 344–345; October 15, 1959.

538.221 3586
Ferromagnetic Domains and Lattice Defects in Cobalt—K. H. von Klitzing and A. Pietzger. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 50–57; July 21, 1959.) Domain patterns obtained with Co powder specimens are discussed.

538.221 3587
Manganese-Arsenic-Antimony Alloys with Invar Characteristics—W. Köster and E. Braun. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 66–68; July 21, 1959.) Investigation of the variation of volume expansion with increasing MnSb content in the mixed-crystal series MnAs-MnSb.

538.221 3588
The Occurrence of Negative Barkhausen Jumps—J. Kranz and A. Schauer. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 84–88; July 21, 1959.) Observations of Weiss domains on the surface of polycrystalline Si-Fe using the magneto-optical Kerr effect [1281 of 1959 (Kranz and Drechsel)] are illustrated and discussed.

538.221 3589
Coupling between Barkhausen Jumps as a Consequence of Magnetic After-Effects—H. Bittel and I. Westerboer. (*Ann. Phys. (Lpz.)*, vol. 4, pp. 203–215; July 21, 1959.) Investigations of the statistical nature of pulse sequences caused by Barkhausen jumps were made with the aid of magnetic tape recordings.

538.221 3590
The Measurement of the Propability Distribution of Barkhausen Jumps in a Ferromagnetic—H. Wilde and H. Girke. (*Z. angew. Phys.*, vol. 11, pp. 339–342; September, 1959.) The probability density distribution of Barkhausen jumps in the Preisach diagram is determined experimentally for a high-permeability ferrite and a perminvar-type ferrite.

538.221:539.23:621.318.57 3591
Nanosecond Switching in Thin Magnetic Films—W. Dietrich, W. E. Proebster, and P.

Wolf. (*IBM J. Res. & Dev.*, vol. 4, pp. 189–196; April, 1960.) Film switching signals as short as 10^{-9}s have been obtained, and thinner films switch faster than thicker ones. The slopes of the inverse-switching-time/driving-field curves are about 10^8 per oersted-second.

538.221:539.234 3592
Measurement of the Magnetic Anisotropy of Vapour-Deposited Thin Ni Films—W. Schüppel and Z. Málek. (*Naturwissenschaften*, vol. 46, pp. 423–424; July, 1959.) Anisotropy measurements were made on films about 1000-Å thick deposited on a glass base at 300° C or at room temperature in a magnetic field of 210 oersted. The hot-deposited films have considerably smaller anisotropy.

538.221:539.234 3593
The Induced Magnetic Anisotropy in Evaporated films of Nickel-Iron Alloys—M. Prutton and E. M. Bradley. (*Proc. Phys. Soc.*, vol. 75, pp. 557–574; April 1, 1960.) A detailed description is given of the experimental technique for preparing Ni-Fe films about 1500-Å thick containing 80–85 per cent Ni. Values of the anisotropy field derived from experimental results are compared with those obtained from a simple model; some agreement is found, but the behavior of the induced anisotropy is shown to be complex, and the possibility of contributions due to imperfection ordering is discussed.

538.221:621.318.134 3594
Crystal Chemical and Magnetic Studies of Garnet Systems $\{YCa_2\}[M_2^{4+}](Fe_3)O_{12}$ $-\{Y_3\}[Fe_2](Fe_3)O_2$, $M = Zr$ **or** Hf—S. Geller, R. M. Bozorth, C. E. Miller, and D. D. Davis. (*J. Phys. Chem. Solids*, vol. 13, pp. 28–32; May, 1960.) See also 2096 of June (Geller, *et al.*).

538.221:621.318.134 3595
Superexchange Interaction in Ferrimagnetic Garnets and Spinels which Contain Randomly Incomplete Linakges—M. A. Gilleo. (*J. Phys. Chem. Solids*, vol. 13, pp. 33–39; May, 1960.)

538.221:621.318.134 3596
Hexagonal Ferrimagnetic Compound containing Fluorine—E. H. Frei, M. Schieber, and S. Shtrikman. (*Phys. Rev.*, vol. 118, p. 657; May 1, 1960.) The compound has a saturation magnetization 7.5 per cent greater than the compound without fluorine.

538.221:621.318.134:538.569.4 3597
Ferromagnetic Resonance in Polycrystalline Ferrites using Circularly Polarized Radiation—R. J. Zeender and E. Schlömann. (*J. Appl. Phys.*, vol. 31, p. 1112; June, 1960.) An investigation of ferromagnetic resonance in the Ni ferrite-aluminate system for compositions near the compensation point is made.

538.221:621.318.134:538.569.4 3598
Ferrimagnetic Resonance in Polycrystalline Europium-Iron Garnet ($Fu_3Fe_5O_{12}$)—Y. Schichijō, T. Miyadai, and H. Takata. (*J. Phys. Soc. Japan*, vol. 15, p. 530; March, 1960.)

538.221:621.318.57 3599
The Different Switching Modes and the Impedance of a Magnetostatic Circuit—C. Durante. (*Compt. rend. acad. sci., Paris*, vol. 249, pp. 2029–2030; November 16, 1959.) An extension of earlier work (253 of January) allowing the same general formulas to be used to determine the behavior of a ferromagnetic core whether the excitation winding is fed from a voltage or a current source. A general expression is given for the impedance of the circuit.

538.222:538.569.4 3600
Spin Lattice Relaxation in Neodymium Ethyl Sulphate at Liquid Helium Temperatures—J. M. Daniels and K. E. Rieckhoff. (*Canad. J. Phys.*, vol. 38, pp. 604–615; May, 1960.) An experimental procedure for measuring relaxation times is described, based on optical Faraday rotation.

538.222:538.569.4 3601
Harmonic Spin Coupling in Ruby—J. E. Geusio (*Phys. Rev.*, vol. 118, pp. 129–130; April 1, 1960.) "A new mode of maser pumping which makes use of harmonic spin coupling in ruby has been demonstrated. In addition, higher-order harmonic spin coupling effects in ruby have been found experimentally."

538.632:537.311.33 3602
The Frequency Characteristics of Hall Generators—F. Kuhrt, H. J. Lippmann, and K. Wiehl. (*Arch. elekt. Übertragung*, vol. 13, pp. 341–347; August, 1959.) Theoretical and experimental investigation of the behavior of Hall generators placed in a high-frequency alternating magnetic field.

621.315.613:538.566.2 3603
RF Electrical Properties of Salty Ice and Frozen Earth—J. C. Cook. (*J. Geophys. Res.*, vol. 65, pp. 1767–1771; June, 1960.) The dielectric constant and resistivity were measured at 100 Mc on artificially produced samples.

MATHEMATICS

512.831 3604
Rational Calculation of Multiple Matrix Products—M. Dreikorn and F. Stockinger. (*Arch. elekt. Übertragung*, vol. 13, pp. 303–306; July, 1959.) A scheme is described for speeding up calculating procedure; results are in the form of a sum of products, which facilitates numerical evaluation by computer.

517.75 3605
Mixed Transcendental and Polynomial Transforms—O. J. M. Smith. (*Commun. and Electronics*, pp. 786–795; January, 1960.) A general exposition of the application of transform techniques includes an assessment of the problem of the direct evaluation of residues and root loci for systems involving feedback around cascaded delay compensators and lumped-parameter elements.

519.15:621.372.54 3606
On a Class of Semi-separable Processes—H. K. Lubbock. (*J. Electronics Control*, vol. 8, pp. 67–79; January, 1960.) The class is defined and its properties are summarized. An example of its application to a filter problem is given.

MEASUREMENTS AND TEST GEAR

529.786+621.3.018.41(083.74) 3607
The International Coordination of Time Signals and Standard-Frequency Transmissions—L. Essen. (*Nature (London)*, vol. 187, pp. 452–453; August 6, 1960.) A review of progress made in U.K. and U.S.A. in the coordination of time signals based on Ephemeris Time and those derived from Cs atomic standards. The feasibility of providing a single service with world-wide coverage is briefly discussed.

529.786+621.3.018.41(083.74) 3608
The Group of Quartz Clocks of the Physikalisch-Technische Bundesanstalt and Comparative Measurement against Atomic Standards—A. Scheibe, U. Adelsberger, G. Becker, G. Ohl, and R. Süss. (*Z. angew. Phys.*, vol. 11, pp.352–357; September, 1959.) Report on the installation and operation of five quartz clocks [see also 3468 of 1956 (Scheibe, *et al.*)]. Frequency differences are plotted for one or more of these clocks and for the NPL Cs resonator (based on MSF transmissions) relative to a reference frequency, over the period 1956.0–1959.0. Fluctuations of the order of a few parts in 10^9 of the NPL standard appear to be confirmed by independent comparisons of the NPL Cs standard with the NBS standard (MSF/WWV) and with the Swiss standard at Neuchâtel [915 of 1959 (Blaser and Bonanomi)].

621.3.018.41(083.74) 3609
A Comparison of Atomic-Beam Frequency Standards—R. E. Beehler, R. C. Mockler, and C. S. Snider. (*Nature (London)*, vol. 187, pp. 681–682; August 20, 1960.) Two dissimilar caesium-beam standards have been compared; their frequencies agree to within ± 1.5 parts in 10^{11}. Effects introduced by the measuring devices themselves which contribute most significantly to the uncertainties of absolute frequency measurement are discussed. These uncertainties have been reduced to a level below 1 part in 10^{11}.

621.3.018.41(083.74) 3610
Measurement Methods and Results for the Standard-Frequency Transmission of the Physikalisch-Technische Bundesanstalt over the Transmitter DCF 77—G. Ohl. (*Nachrichtentech. Z.*, vol. 12, pp. 374–375; July, 1959.) Methods of frequency measurement by comparison with a crystal clock or other standard-frequency transmissions are described. The curve of frequency stability of Station DCF 77 [2857 of August (Adelsberger)] is given for March, 1959.

621.317.3:621.315.212 3611
Shielded Coaxial Leads for Low-Temperature Electrical Measurements—N. L. Brown and R. N. Barfield. (*Rev. Sci. Instr.*, vol. 31, pp. 517–519; May, 1960.) Doubly shielded rigid coaxial leads are described for use in dielectric measurements down to 4.2°K. They have low heat losses and small and stable interelectrode capacitances.

621.317.33:537.311.33 3612
High-Frequency Measurement of the Resistivity of Semiconductor Crystals—W. Keller. (*Z. angew. Phys.*, vol. 11, pp. 346–350; September, 1959.) Several methods of contactless HF measurement are summarized, and details are given of a tuned-circuit method with capacitive coupling to the semiconductor rod, in this case single-crystal Si. See also 3559 above.

621.317.332.1+621.317.361]:621.396.663 3613
Goniometer Measuring Systems for High Frequencies: Part 1—Goniometer Design, Admittance Measurement, Frequency Measurement—H. Fricke. (*Arch. tech. Messen*, No. 283, pp. 153–156; August, 1959.)

621.317.335.3.029.6:621.372.826 3614
Application of Slow Surface Waves to the Measurement of Dielectric Constants of Substances at Ultra High Frequencies—V. P. Shestopalov and K. P. Yatsuk. (*Radiotekhnika i Elektronika*, vol. 4, pp. 547–549; March, 1959.) Note of a method based on wavelength measurement in a helix with and without the dielectric.

621.317.39:531.781.2 3615
Oscillator Measuring Equipment for Vibrating-Wire Gauges—W. H. Ward and J. E. Cheney. (*J. Sci. Instr.*, vol. 37, pp. 88–92; March, 1960.) Frequencies of vibrating-wire gauges are measured by comparison with a variable-frequency tube oscillator which is itself checked against a tuning fork and time signals. The instrument covers a range of 250–2500 cps and is cheaper and simpler to operate than traditional types .

621.317.6:621.374.32 3616
Digital Measurement of Axis-Crossing Intervals—A. J. Rainal. (*Electronics*, vol. 33, pp.

88–91; June 3, 1960.) Description of an instrument for statistical sampling with applications in the study of weak-signal detection and noise.

621.317.7:621.318.57:539.23 3617
Measuring Switching Speed of Magnetic Films—W. Dietrich and W. E. Proebster. (*Electronics*, vol. 33, pp. 79–81; June 3, 1960.) A charged coaxial cable terminated in a shorted strip line produces a fast-rise pulse for measuring polarity reversal time of thin magnetic films.

621.317.71:538.632 3618
Direct Measurement of $\int i^2 \cdot dt$ by a Method based on the Hall Effect—S. C. Ryder-Smith, A. E. Guile, and A. E. Barrington. (*J. Sci. Instr.*, vol. 37, pp. 100–103; March, 1960.) Details are given of a squaring unit comprising two InSb crystals which gives an ac output proportional to the square of the input current. The output after being amplified and rectified is fed to a fluxmeter. The instrument has been developed to investigate the characteristics of fuses and some results are given.

621.317.733.011.21 3619
AC Bridges with Inductive Ratio Arms—F. Butler. (*Electronic Tech.*, vol. 37, pp. 303–309; August, 1960.) Errors occurring with transformers of normal design may be greatly reduced by a procedure involving a preliminary balance followed by a second balance with the unknown impedance short-circuited, or by applying a calculable correction.

621.317.733.011.4 3620
Using Off-Balance Bridges for Measurement and Control—G. Revesz. (*Electronics*, vol. 33, pp. 52–54; July 1, 1960.) Theory and design details of an unbalanced transformer bridge for the measurement of small capacitances are given, and methods of reducing the inherent nonlinearity are discussed.

621.317.755 3621
Converting Oscilloscopes for Fast Rise Time Sampling—J. J. Amodei. (*Electronics*, vol. 33, pp. 96–99; June 24, 1960.) A sampling technique, using transistors in the avalanche mode, can resolve rise times of $\frac{1}{3}$ mμsec, using a conventional 10-Mc oscilloscope. Sensitivities between 10 and 30 mv/cm are obtainable. A full circuit diagram is included.

621.317.761.083.7:621.396.934 3622
F.R.E.S.C.O.—an Airborne Frequency Meter with Digital Output—J. Ackroyd. (*Brit. Commun. Electronics*, vol. 7, pp. 502–507; July, 1960.) In the equipment described, the output frequency of a pressure transducer used for pressure measurements in missiles is automatically measured and the result coded in a form suitable for transmission by a telemetry system.

621.317.789.029.6 3623
Microwave Attenuation Measurements with Accuracies from 0.0001 to 0.06 Decibel over a Range of 0.01 to 50 Decibels—G. F. Engen and R. W. Beatty. (*J. Res. NBS*, vol. 64C, pp. 139–145; April–June, 1960.) The theory of measurement is discussed and block diagrams are given of the equipment used. The results obtained in calibrating a rotary-vane type of attenuator are evaluated.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.362+621.56 3624
Design Parameters for Optimizing the Efficiency of Thermoelectric Generators Utilizing *p*-Type and *n*-Type Lead Telluride—R. W. Fritts. (*Commun. and Electronics*, pp. 817–820; January, 1960.) Results are presented which indicate the variation with temperature of electrical resistivity, thermal conductivity and Seebeck coefficient. It is shown that optimum design requires that the crystal should be doped to provide maximum figures of merit at given intervals in the temperature gradient.

621.365.55:621.373.421.14.029.6 3625
Parallel Operation of Several Input Coupling Systems in Microwave Generators with Enclosed Work Chambers—W. Schmidt. *Elektronische Rundschau*, vol. 13, pp. 280–282; August, 1959.) Parallel operation of magnetrons with independent feed systems is considered; suitable phasing of the inputs smoothes the energy flow distribution. See 619 of February.

621.385.833 3626
Numerical Investigation of a Typical Electrostatic Immersion Objective—E. Hahn. (*Optik, Stuttgart*, vol. 16, pp. 513–521; September, 1959.)

621.387:621.372.413 3627
Microwave Discharge Chamber—S. Fukui and S. Hayakawa. (*J. Phys. Soc. Japan*, vol. 15, p. 532; March, 1960.) The advantages of a microwave cavity as a discharge chamber for particle detection are discussed.

621.387:621.374.3 3628
A 2000-Channel Analyser for Neutron Spectroscopy—J. Hahn and W. W. Havens, Jr. (*Rev. Sci. Instr.*, vol. 31, pp. 490–500; May, 1950.) Neutron times-of-flight are recorded on a fast electrostatic store and then transferred to a large-capacity drum store between the neutron pulses. Full details are given of all circuitry, including test equipment.

621.398 3629
Transportable Ground Receiving and Recording Equipment for 24-Channel Telemetry—F. G. Diver. (*J. Brit. IRE*, vol. 20, pp. 457–464; June, 1960.) A carrier in the 400–500-Mc band is amplitude modulated by a FM subcarrier using time-division multiplexing. Output data are available in "histogram" form or as individually selected channel groups.

621.398:621.376.55 3630
A 600-Mc Pulse Telemetry System—G. T. Hardwick. (*J. Brit. IRE*, vol. 20, pp. 465–473; June, 1960.) A brief account of the airborne equipment is given, and the receiving, testing and calibration equipment are described in detail. Information is conveyed by the time delays between a wide datum pulse and up to 20-narrow pulses.

PROPAGATION OF WAVES

621.391.812.6.029.4 3631
Terrestrial Propagation of Very-Low-Frequency Radio Waves: a Theoretical Investigation—J. R. Wait. (*J. Res. NBS*, vol. 64D, pp. 153–204; March/April, 1960.) A treatment of the propagation of TLF waves in terms of waveguide mode theory, taking account of earth curvature as well as the influence of magnetic field, antipodal effects, resonator-type oscillations, and stratification at the lower edge of the ionosphere. Over 80 references.

621.391.812.6.029.4 3632
Influence of Source Distance on the Impedance Characteristics of E.L.F. Radio Waves—J. R. Wait. (Proc. IRE, vol. 48, pp. 1338–1339; July, 1960.) The importance of the wave impedance in the propagation of waves in the frequency range below 3 kc is discussed, and graphs show the variation of wave impedance with source distance.

621.391.812.6.029.55/.62 3633
Peculiarities and Seasonal Variations of Transequatorial Back-Scatter Echoes as Observed at Mayagüez, Puerto Rico—B. Dueño. (*J. Geophys. Res.*, vol. 65, pp. 1691–1698; June, 1960.) Long-range transequatorial echoes (LRTE) between 20 and 50 Mc have a maximum rate of occurrence at the equinoxes and a minimum near the solstices. It is possible that LRTE are the result of large concentrations of electrons near the equator, particularly during the evening. The angle of arrival of LRTE is low, and only a small fraction of the trajectory of the wave lies in the ionosphere.

621.391.812.624 3634
Application of the Turbulence Theory to the Scattering of Radio Waves on Randomly Moving Inhomogeneities—M. I. Rodak and A. V. Frantsesson. (*Radiotekhnika i Elektronika*, vol. 4, pp. 398–403; March, 1959.) It is shown under which conditions it is possible to calculate the corellation function of the scattered field and the intensity assuming that the velocity of the scattering centers is constant.

621.391.812.63 3635
Effects of Nonlinearity on Propagation in Ionized Media—G. I. Cohn. (*Commun. and Electronics*, pp. 942–949; January, 1960.) A theoretical investigation of the interference signals generated when a plane polarized wave is propagated through a medium in which the conductivity is nonlinear.

621.391.812.63 3636
On the Absorption of Radio Waves in Resonant Regions of Heterogeneous Plasma—N. G. Denisov. (*Radiotekhnika i Elektronika*, vol. 4, pp. 388–397; March, 1959.) Investigation of the passage of waves through a medium having a refractive index which tends to infinity. In this region, the frequency of the external field coincides with the natural frequency of the plasma oscillations. The effect of wave absorption in the resonance region is calculated neglecting interaction. The results are considered in connection with wave propagation in the ionosphere.

621.391.812.63 3637
Polarization Parameters of the Downcoming Radio Waves—Y. S. N. Murty and S. R. Khastgir. (*J. Geophys. Res.*, vol. 65, pp. 1449–1457; May, 1960.) Discrepancies in the results obtained for the polarization parameters by various workers are only apparent, and arise out of differences of form and notation in the formulas used.

621.391.812.63.029.45:551.510.535 3638
V.L.F. Propagation Effects of a D-Region Layer Produced by Cosmic Rays—W. F. Moler. (*J. Geophys. Res.*, vol. 65, pp. 1459–1468; May, 1900.) A model is proposed in which a cosmic-ray layer $D\beta$ is produced below the photoionization layer $D\alpha$, the height and intensity of $D\beta$ being functions of geomagnetic latitude. Calculated phase and amplitude characteristics of VLF signals reflected from the $D\beta$ layer during the presunrise hour are in good agreement with those measured.

621.391.812.63.029.62 3639
Refraction of V.H.F. Signals at Ionospheric Heights—S. Weisbrod and L. Colin. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 107–109; January, 1960.) Unlike tropospheric refraction, which decreases rapidly with elevation angle, refraction in the ionosphere reaches a maximum value for angles of about 0.1 radius. A graph showing the variation of deviation with elevation angle for both the ionosphere and troposphere is given.

621.391.812.63.029.62 3640
Radio Echoes from Field-Aligned Ionization above the Magnetic Equator and their Resemblance to Auroral Echoes—K. L. Bowles, R. Cohen, G. R. Ochs, and B. B. Balsley. (*J. Geophys. Res.*, vol. 65, pp. 1853–1855; June, 1960.) Doppler shift measurements are inconsistent with a simple drift motion of irregularities. It is, therefore, suggested that both equatorial and auroral irregularities are aligned in the local field which, at least for equatorial regions, is seriously distorted by the flow of the electrojet.

621.391.812.63.029.63 3641
Ionospheric Back-Scatter Observations at 440 Mc—V. C. Pineo, L. G. Kraft and H. W. Briscoe. (*J. Geophys. Res.*, vol. 65, pp. 1620–1621; May, 1960.) Back scatter signals from heights of 100–800 km have been observed in Massachusetts at elevation angles down to 15° under winter afternoon conditions. Vertical-incidence frequency-spectrum measurements in a band 30 kc wide centered on 440 Mc have also been made.

RECEPTION

621.391.812.36.029.63 3642
Received Amplitude and Fading Distribution at 10-cm Wavelength as a Function of Azimuthal Deviation from the Great Circle—R. Schünemann and G. Pucher. (*Hochfrequenztech. u. Elektroakust.*, vol. 68, pp. 37–42; July, 1959.) Report of measurements over a 76-km path made with 4-m parabolic reflectors at 10.2 cm λ. Results appear to favor a model atmosphere that gives rise to partial reflections rather than scattering.

621.391.812.6 3643
The Diurnal Variations of the Back-Scatter Field Strength of Telegraphy Signals—B. Beckmann and K. Vogt. (*Nachrichtentech. Z.*, vol. 12, pp. 419–420; August, 1959.) The use of backscatter measurements by means of the "ring" method (1888 of 1957) for determining propagation conditions is exemplified by a film record of diurnal variations of back scatter at a frequency of 25.42 Mc in the direction towards North America. This is related to simultaneous field-strength measurements of Station WWV also recorded in West Germany.

621.391.82 3644
The External Noise of Commercial Aerial Installations—W. Kronjäger and K. Vogt. (*Nachrichtentech. Z.*, vol. 12, pp. 371–373; July, 1969.) Measurements were made of frequency distribution in the range 2–30 Mc of the external noise received by the rhombic antennas at the receiving stations Eschborn [1634 of 1958 (Kronjäger, *et al.*)] and Lüchow. Pronounced differences in noise reception at frequencies below about 8 Mc are caused by the proximity of sources of man-made noise to the Eschborn station.

621.391.82:523.164 3645
Extraterrestrial Radio Noise as a Source of Interference in the Frequency Range 30–1000 Mc—F. Horner. (*Proc. IEE*, vol. 107, pt. B, pp. 373–376; July, 1960.) "Published information on the intensity of noise from the galaxy, the sun, and from other extraterrestrial sources is presented in a form which shows their importance relative to the internal noise of typical receiving installations. Particular attention is paid to a half-wave horizontal dipole antenna and to an antenna with a 20° pencil beam directed horizontally."

STATIONS AND COMMUNICATION SYSTEMS

621.394:621.391.812.62.029.63 3646
Measured Statistical Characteristics and Narrow-Band Teletype Message Errors on a Single-Sideband 600-Mile-Long Ultra-High-Frequency Tropospheric Radio Link—E. F. Florman and R. W. Plush. (*J. Res. NBS.*, vol. 64D, pp. 125–133; March/April, 1960.) A description of tests made on a 417-Mc unmodulated carrier, using antenna beamwidths of about 1°.

621.396.41 3647
Asynchronous Multiplexing—J. E. Taylor. (*Commun. and Electronics*, No. 46, pp. 1054–1062; January, 1960.) Coding theory is applied to the evaluation of the optimum information rate and the number of channels available.

621.396.46 3648
Optimum Frequencies for Outer Space Communication—G. W. Haydon. (*J. Res. NBS*, vol. 64D, pp. 105–109; March/April 1960.) A discussion of the problems in communication between earth and a space vehicle, and between space vehicles. Special reference is made to the band 10–10000 Mc.

SUBSIDIARY APPARATUS

621–52 3649
On the Stability of Random Systems and the Stabilization of Deterministic Systems with Random Noise—J. C. Samuels. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 594–601; May, 1960.) A general theory of mean square stability of random linear systems in which parameters may vary as white noise processes is developed and applied to a RLC circuit. The possibility of stabilizing unstable deterministic systems by means of random noise is predicted.

621.3.087.4:621.319.4 3650
A Condenser Memory Unit for Improving Signal-to-Noise Ratios—D. M. Hunten. (*Canad. J. Phys.*, vol. 38, pp. 346–353; March, 1960.) Description of a unit containing 30 low-leakage capacitors which can store information for periods up to several hours; 32 channels may be scanned in 10 seconds. The unit has been used in investigations of twilight spectra with photoelectric and photoconductive spectrometers.

621.3.087.6:621.375.4 3651
A Sensitive Transistor Recording-Pen Amplifier—K. G. Beauchamp. (*Electronic Engrg*, vol. 32, pp. 444–445; July, 1960.) Details of a bridge circuit with low temperature drift are given.

621.311.6:621.382.3 3652
A Magnetic-Amplifier/Silicon-Transistor Power Supply for Missile Application—B. Mokrytzki and R. A. Stuart. (*Commun. and Electronics*, No. 46, pp. 835–843; January, 1960.) A detailed description of a power supply system producing seven regulated direct voltages and two rms-regulated alternating voltages. A magnetic amplifier is incorporated in each channel for regulation and all direct voltages are derived from the same HF transistor inverter.

621.311.62.078:621.318.381 3653
A Stabilized and Programme-Controlled Power Supply using Positive Feedback—M. Pacāk. (*Electronic Engrg.*, vol. 32, pp. 372–373; June, 1960.) A power supply for a mass-spectrometer magnet is described with an output continuously variable from 3 to 800 volts and a short-term stability better than 20 mv.

621.311.69:535.215 3654
Limitations and Possibilities for Improvement of Photovoltaic Solar Energy Converters: Part 1—Considerations for Earth's Surface Operation—M. Wolf. (Proc. IRE, vol. 48, pp. 1246–1263; July, 1960.) The possibilities of improving the performance of the single Si solar cell are considered; as these are limited, the use of other materials and the development of a multilayer and a multiple transition cell are examined, and found to give larger theoretical efficiencies.

621.311.69:621.396.65 3655
The Wind-Driven Power Supply for the Decimetre-Wave Radio Relay Station Schöneberg (Eifel) and the Experience Gained with it—G. Rösseler. (*Nachrichtentech. Z.*, vol. 12, pp. 352–360; July, 1959.) Details are given of the wind-driven generators and ancillary equipment used in a 22-month trial of the system as the main power supply for radio-link stations. Records of wind-power availability and utilization made by specially designed apparatus are analyzed and discussed.

621.314.63:546.289 3656
The Parallel and Series Operation of Germanium Rectifiers—H. S. Blanks. (*Proc. IRE, (Australia)*, vol. 20, pp. 406 412; July, 1959.) "The permissible ratings of germanium rectifiers operated in series and parallel connections have been calculated for the cases where load balancing resistors are placed in parallel and in series, respectively, with the individual rectifiers."

621.316.72:621.318.381:621.382.3 3657
Transistorized Power Supplies for a Mass Spectrometer—R. D. Russell and F. Kollar. (*Canad. J. Phys.*, vol. 38, pp. 616–623; May, 1960.) Magnet current supply and filament emission control circuits of low power dissipation are described.

TELEVISION AND PHOTOTELEGRAPHY

621.397.132 3658
Investigation of the Fine Structure of Colour Television Images—I. Bornemann. (*Nachrtech.*, vol. 9, pp. 313–317; July, 1959.) The formation of color images consisting of a mosaic of dots in three primary colors is considered, and the information entropy of the image is evaluated on the basis of dot size and brightness. The resolving power of shadow-mask tricolor picture tubes is also discussed.

621.397.132 3659
The Practical Effect of Frequency-Band Limitation on Colour Television Images—P. Neidhardt. (*Nachrtech.*, vol. 9, pp. 318–323; July, 1959.) The transient response of linear ($\gamma = 1$) and nonlinear NTSC color-television systems is investigated as a function of bandwidth and the effects on color reproduction are analyzed. 24 references.

621.397.132 3660
Henri de France Colour Television System—(*Electronic Engrg.*, vol. 32, pp. 356–358; June, 1960.) Chrominance information is sequentially transmitted in alternate lines and is stored at the receiver from line to line. Two signals are transmitted simultaneously instead of three, and the chrominance subcarrier is modulated by a single signal.

621.397.132:389.6(4) 3661
Standardization Problems of European Colour Television—W. Gerber. (*Tech. Mitt. PTT*, vol. 37, pp. 262–267; July 1, 1959.) Review of color television systems and discussion of their adaptability to existing European standards. A modified 625-line CCIR standard with 8-Mc channel width and 5.5-Mc sound-vision carrier separation may offer the best solution.

621.397.331.21 3662
A New Method for Evaluating the Resolving Power of Television Camera Tubes—Ya. A. Ryftin and M. V. Antipin. (*Zh. Tekh. Fiz.*, vol. 29, pp. 252–260; February, 1959.) The method is based on measuring the number of "normal"

scanning lines, referred to a square scanning spot, by means of a test card with a pattern of Fresnel regions.

621.397.621 3663
Two-Oscillator Scan Stability—H. M. Joseph. (*Commun. and Electronics*, No. 46, pp. 1022–1028; January, 1960.) Formulas are developed for choosing the frequency ratio and the relative stability required when two separate oscillators are used to generate a television raster.

621.397.7:621.315.212 3664
A New Method for the Suppression of Low-Frequency Noise on Television Cables—R. Rasch. (*Nachrichtentech. Z.*, vol. 12, pp. 416–418; August, 1959.) Description of a specially developed unit for reducing the mains hum on coaxial-cable links and for improving the signal-to-noise ratio at low frequencies for crosstalk on parallel video transmission lines.

621.397.7:621.373.43.072.9 3665
The Synchronization of Television Pulse-Waveform Control Centres—H. Windischbauer. (*Elektronik*, vol. 18, pp. 233–237; August, 1959.) Various techniques of synchronizing the sources of sync pulses are described with block diagrams.

621.397.743 3666
Sync-Signal Regeneration System for Low-Power Translators—P. K. Kirilloff. (*J. Brit. IRE*, vol. 20, pp. 475–477; June, 1960.) A noise-immune synchronization system is discussed, and a brief description given of the basic filter circuits.

621.397.9:522.2 3667
Television System for Stratoscope I—L. E. Flory, G. W. Gray, J. M. Morgan, and W. S. Pike. (*Electronics*, vol. 33, pp. 49–53; June 17, 1960.) A detailed description is given of a television telemetry system used with a balloon-borne optical telescope.

621.397.9(47) 3668
Application of Television in Industry and Science in the U.S.S.R.—V. I. Sardyko. (*J. Brit. IRE*, vol. 20, pp. 449–456; June, 1960.) Television equipment developed for nondomestic use with applications in medicine, metallurgy and transport is described with an outline of future development.

TUBES AND THERMIONICS

621.382 3669
Negative Resistance in Semiconductor Devices—R. E. Burgess. (*Canad. J. Phys.*, vol. 38, pp. 369–375; March, 1960.) The ac impedance characteristics arising from the interaction of thermal and nonthermal effects are discussed, and criteria for the appearance of negative resistance over certain ranges of frequency are derived. The application of this analysis to devices which do and do not exhibit isothermal negative resistance is considered.

621.382.22 3670
The Effective Contact Area of Point-Contact Crystal Rectifiers—E. Groschwitz and R. Ebhardt. (*Z. angew. Phys.*, vol. 11, pp. 342–346; September, 1959.) The mechanism of rectification is discussed quantitatively on the basis of earlier work (3546 above) and taking account of the presence of a surface current in an inversion layer under low-voltage conditions.

621.382.23 3671
Biasing Methods for Tunnel Diodes—R. P. Murray (*Electronics*, vol. 33, pp. 82–83; June 3, 1960.) A description of methods for determining the correct bias and circuit impedance conditions for operation of a tunnel diode as an oscillator, amplifier, or switch.

621.382.23 3672
Tunnel Diodes by Vapour Growth of Ge on Ge and on GaAs—J. C. Marinace. (*IBM J. Res. & Dev.*, vol. 4, pp. 280 282; July, 1960.) Degenerately doped n-type Ge can be vapor-grown epitaxially on degenerately doped p-type Ge or GaAs substrates, yielding tunnel junctions in either case.

621.382.23 3673
Electrical Properties of Vapour-Grown Ge Junctions—M. J. O'Rourke, J. C. Marinace, R. L. Anderson, and W. H. White. (*IBM J. Res. & Dev.*, vol. 4, pp. 256–263; July, 1960.) Characteristics of several different kinds of p-n junction fabricated by a vapor growth process (see 3563 above) are given. Ge diodes produced by this process compare favorably with those produced by other means.

621.382.23 3674
A Vapour-Grown Variable-Capacitance Diode—R. L. Anderson and M. J. O'Rourke. (*IBM J. Res. & Dev.*, vol. 4, pp. 264–268; July, 1960.) Capacitance/voltage characteristics are discussed in terms of impurity distributions.

621.382.23:621.375.9 3675
Diode Capacitors for Parametric Amplification—R. C. Knechtli and R. D. Weglein: K. E. Mortenson. (*J. Appl. Phys.*, vol. 31, pp. 1134–1135; June, 1960.) Comment on 441 of February and author's reply.

621.382.3.012.8 3676
The High-Frequency Equivalent Circuit of the Transistor—R. Paul. (*Nachrtech.*, vol. 9, pp. 296–300; July, 1959.) A simplified equivalent circuit is derived for the earthed-base transistor.

621.382.333 3677
Avalanche Carrier Multiplication in Junction Transistors and its Implications in Circuit Design—M. L. N. Forrest. (*J. Brit. IRE*, vol. 20, pp. 429–439; June, 1960.) Theories of junction breakdown and transistor action necessary to account for the observed characteristics are surveyed. Design techniques are discussed and methods of avoiding anomalous operation are given.

621.382.333 3678
Some Interesting Transistor Characteristics in the Millimicroampere Region—R. A. Dandl and F. T. May. (*Rev. Sci. Instr.*, vol. 31, pp. 575–576; May, 1960.) Measurements are given showing that some transistors can have very high current gains when the base current is about 10^{-10}A.

621.382.333:539.12.04 3679
On the Neutron-Bombardment Reduction of Transistor Current Gain—J. W. Easley and J. A. Dooley. (*J. Appl. Phys.*, vol. 31, pp. 1024–1028; June, 1960.) Measurements of the decay of Ge alloy transistor current gain caused by fast-neutron and gamma-ray bombardment show that n-p-n types give a linear dependance of current gain on exposure, but p-n-p types do not generally do so. The factors leading to this behavior are discussed.

621.382.333:621.391.827 3680
Investigation of the Cross-Modulation Characteristics of H. F. Transistors—H. Lotsch. (*Elektronische Rundschau*, vol. 13, pp. 290–294; August, 1959.) The cross-modulation properties affecting the reception of a weak signal are calculated from the transistor characteristics and from the third-harmonic distortion factor, and the results are compared with those of HF and LF measurements.

621.382.333.33 3681
Minority Carrier Recombination in a Cylindrical Transistor Base Region—D. P. Kennedy. (*J. Appl. Phys.*, vol. 31, pp. 954–956; June, 1960.) "An analysis is given on the influence of bulk recombination within the base region of a mesa-type drift transistor. The minority carrier transport efficiency is established for a solid cylinder base region and also for a simplified one-dimensional structure. A comparison of the two minority carrier transport equations shows the approximate analysis will result in a negligible error when applied to practical semiconductor devices."

621.383.81 3682
Image Intensification by Transmission Secondary Electron Emission—M. M. Wachtel, D. D. Doughty, G. Goetze, A. E. Anderson, and E. J. Sternglass. (*Rev. Sci. Instr.*, vol. 131, pp. 576–578; May, 1960.) Amplified images with good resolution are obtained with a multistage intensifier using a series of plane-parallel thin-film dynodes of BaF_2 to multiply electrons from a Cs_3Sb photocathode.

621.385.2:621.362 3683
On the Theory of the Close-Spaced Impregnated-Cathode Thermionic Converter—E. S. Rittner. (*J. Appl. Phys.*, vol. 31, pp. 1065–1071; June, 1960.) Formulas which approximate very closely to the exact Langmuir space-charge theory are used, with a computer, to study the performance of a thermionic converter comprising two closely spaced planar impregnated cathodes.

621.385.2.029.6:621.391.822.3 3684
Transport of Noise at Microwave Frequencies through a Space-Charge-Limited Diode—W. E. Vivian. (*J. Appl. Phys.*, vol. 31, pp. 957–962; June, 1960.) A new method of analysis is outlined which eliminates inconsistencies in previous work based on a (perturbed) monovelocity model, and permits resolution of the infinite transit time problem. The magnitude and variation with distance of the "beam noise invariants" is calculated for a range of diode operating conditions, and shown to be consistent with experimental data.

621.385.623.5 3685
Determination of the Coefficient of Phase Aberration Φ in a Reflex Klystron—J. Bonnerot. (*Compt. rend. acad. sci., Paris*, vol. 249, pp. 2023–2025; November 16, 1959.) Φ, the ratio of the actual beam current to the optimum calculated current was found to be 0.36.

621.385.623.5:621.376.23 3686
Detection in the Cathode Circuit of an Underexcited Reflex Klystron—S. A. Kornilov and O. N. Kazbekova. (*Radiotekhnika i Elektronika*, vol. 4, pp. 475–381; March, 1959.) An experimental investigation of the detection characteristics, the sensitivity and the transfer coefficient of the cathode circuit in a klystron amplifier, shows that these characteristics are not inferior to those of a crystal detector.

621.385.623.5:376.23 3687
Application of Virtual Cathodes for Detection at Ultra High Frequency—A. I. Kostienko, M. N. Devyatkov, and A. A. Lebed'. (*Radiotekhnika i Elektronika*, vol. 4, pp. 482–488; March, 1959.) An experimental investigation using a reflex klystron at wavelengths of 10 and 3 cm. Results are shown graphically.

621.385.63:621.375.9:621.372.44 3688
Space-Charge Wave Parametric Amplifiers—J. S. Cook, W. H. Louisell, and C. F. Quate. (*J. Electronics Control*, vol. 8, pp. 1–18; January, 1960.) A detailed theoretical examination suggests that noise caused by high idler frequencies may be avoided by coupling the sig-

nal and pump to the beam by means of a slow-wave structure. Adjustment of the phase velocities of the pump and signal ensures that eventually the noise is transferred to the beam and the signal to the slow-wave structure.

621.385.63:621.396.65 3689
The Limits of Amplification of Travelling-Wave Valves for Radio Links—W. Klein. (*Arch. elekt. Übertragung*, vol. 13, pp. 273–286; July, 1959.) The parameters affecting tube gain are discussed, and the limits set by noise and self-oscillation caused by internal reflections are investigated. The internal reflections are calculated for a nonuniform helix with localized attenuation; measurements confirm the results of the theoretical investigations.

621.385.632.14 3690
Travelling-Wave Amplification of a Spatial Harmonic Tube using a Comb-Type Structure—Y. Kato and T. Kobayashi. (*Rep. Elect. Commun. Lab., Japan*, vol. 7, pp. 321–332; September, 1959.) Experimental results obtained with a 4-kMc scaled model and with demountable tubes designed for the 23-kMc region are given.

621.385.632.14 3691
A Comb-Structure Delay System for Travelling-Wave Amplifiers—A. M. Belyantsev, A. V. Gaponov, and E. V. Zagryadskiĭ. (*Radiotekhnika i Elektronika*, vol. 4, pp. 504–516; March, 1959.) A method is described in which an equivalent system of line segments is adopted. This reduces the problem to the examination of waves in a multipole circuit with cross connections. Results obtained with different comb structures are shown graphically.

621.385.632.14 3692
Anisotropic Properties of a Corrugated Delay Structure—L. S. Benenson. (*Radiotekhnika i Elektronika*, vol. 4, pp. 517–520; March, 1959.) A formula is derived for the reduction of the phase velocity of surface waves propagated at an arbitrary angle to the corrugated structure. An expression is also obtained, relating the beam direction to the direction of the wave front.

621.385.632.14:621.372.853.1 3693
Electron Flow in a Waveguide Partially Filled with a Dielectric—E. I. Vasil'ev. (*Radiotekhnika i Elektronika*, vol. 4, pp. 500–504; March, 1959.) A dispersion equation is given for calculating propagation constant of circular waveguide with an electron stream traveling along the axis.

621.385.632.3 3694
Nonlinear Theory of Double-Beam Valves: Part 1—Conclusions and Investigation of Equations—G. F. Filimonov. (*Radiotekhnika i Elektronika*, vol. 4, pp. 489–499; March, 1959.)

621.385.633 3695
A Backward-Wave Oscillator with Periodic Delay Line for the Frequency Range 27–48 kMc—F. Gross. (*Arch. elekt. Übertragung*, vol. 13, pp. 356–362; August, 1959.) A backward-wave traveling-wave tube is described rated for a CW output >40 mW. The slow-wave structure consists of a stacked assembly of Cu or Mo disks.

621.385.633.12 3696
Helix-Type Backward-Wave Oscillator (ECL-1182)—M. Kamihara. (*Rep. Elect. Commun. Lab., Japan*, vol. 7, pp. 393–407; November, 1959.) Backward-wave oscillator theory and design procedures are discussed and applied to the development of a medium-power tube in which the beam is focused by a cylindrical permanent magnet.

621.385.64 3697
The Electronic Admittance of a C. W. Magnetron Oscillator—C. G. Lehr and A. J. Kalnajs. (*J. Electronics Control*, vol. 8, pp. 19–37; January, 1960.) An accurate method of measurement is described which shows that the electronic susceptance of a particular magnetron is always positive and that it varies strongly with RF voltage but not with anode current.

621.385.644.2 3698
Calculation of Operating Characteristics of Multicavity Magnetrons—S. I. Bychkov. (*Radiotekhnika i Elektronika*, vol. 4, pp. 468–474; March, 1959.)

621.385.832 3699
A Simple Multidimensional C.R.T. Display Unit—D. M. MacKay. (*Electronic Engrg.*, vol. 32, pp. 344–347; June, 1960.) A simplified version of an earlier circuit (3305 of 1949) is described, using linear instead of sine-cosine potentiometers.

621.385.832.001.4 3700
Astigmatism in Cathode-Ray Tubes—N. Patla. (*J. Inst. Telecommun. Engrs., India*, vol. 6, pp. 31–37; December, 1959.) Details are given of electronic apparatus suitable for mass-production testing of CR tubes. The difference in the value of focusing voltages necessary to achieve the finest definition of the width of a circular pulsating trace in its XX′ and YY′ directions is a measure of the degree of astigmatism.

621.385.832.032.265.2 3701
Electrostatic Deflection Plates for Cathode-Ray Tubes—L. Frenkel. (*J. Res. NBS*, vol. 64C, pp. 103–113; April–June, 1960.)

Part 1—Design of Single-Bend Deflection Plates with Parallel Entrance Sections.—Sets of design curves based on the characteristic peak-to-peak plate cutoff voltage are given for determining the plate dimensions for a given sensitivity and scan.

Part 2—Deflection Defocusing Distortion of Single-Bend Deflection Plates with Parallel Entrance Sections.—A convenient formula for calculating defocusing distortion is derived and compared with experiment.

621.387:621.318.57 3702
General Aspects of Electromagnetic Discharge Phenomena in Gases—L. Goldstein and Musson-Genon. (*Le Vide*, vol. 14, pp. 327–343; November/December, 1959.) Properties of ionized gases are examined and their application is described with reference to the design of different TR cells for microwave radar systems.

621.387:621.318.57 3703
Elimination of Charges in Switching Tubes—J. Bricon. (*Le Vide*, vol. 14, pp. 357–364; November/December, 1959.) In French and English.) Recombination, attachment and diffusion processes for the deionization of TR cells are described. The most satisfactory process is one in which diffusion takes place on quartz fibres in an argon-filled cell.

Translations of Russian Technical Literature

Listed below is information on Russian technical literature in electronics and allied fields which is available in the U. S. in the English language. Further inquiries should be directed to the sources listed. In addition, general information on translation programs in the U. S. may be obtained from the Office of Science Information Service, National Science Foundation, Washington 25, D. C., and from the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C.

PUBLICATION	FREQUENCY	DESCRIPTION	SPONSOR	ORDER FROM:
Acoustics Journal (Akusticheskii Zhurnal)	Quarterly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Automation and Remote Control (Avtomatika i Telemekhanika)	Monthly	Complete journal	National Science Foundation—MIT	Instrument Society of America 313 Sixth Ave., Pittsburgh 22, Pa.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Journal of Abstracts, Electrical Engineering (Reserativnyy Zhurnal: Electronika)	Monthly	Abstracts of Russian and non-Russian literature		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Journal of Experimental and Theoretical Physics (Zhurnal Eksperimentalnoi i Teoreticheskoi Fiziki)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Journal of Technical Physics (Zhurnal Tekhnicheskoi Fiziki)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Proceedings of the USSR Academy of Sciences: Applied Physics Section (Doklady Akademii Nauk SSSR: Otdel Prikladnoi Fiziki)	Bimonthly	Complete journal		Consultants Bureau, Inc. 227 W. 17 St., New York 22, N. Y.
Radio Engineering (Radiotekhnika)	Monthly	Complete journal	National Science Foundation—MIT	Pergamon Institute 122 E. 55 St., New York 22, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Radio Engineering and Electronics (Radiotekhnika i Elektronika)	Monthly	Complete journal	National Science Foundation—MIT	Pergamon Institute 122 E. 55 St., New York 22, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Solid State Physics (Fizika Tverdogo Tela)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Telecommunications (Elekprosviaz')	Monthly	Complete journal	National Science Foundation—MIT	Pergamon Institute 122 E. 55 St., New York 22, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Automation Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
Electronics Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
Physics Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
Express Contents of Soviet Journals Currently being Translated into English	Monthly	Advance tables of contents of translated journals		Consultants Bureau, Inc. 227 W. 17 St., New York 22, N. Y.
Technical Translations	Twice a month	Central directory in the U. S. of translations available from all major sources in the U. S.	OTS and Special Libraries Assoc.	Superintendent of Documents U. S. Gov't Printing Office Washington 25, D. C.

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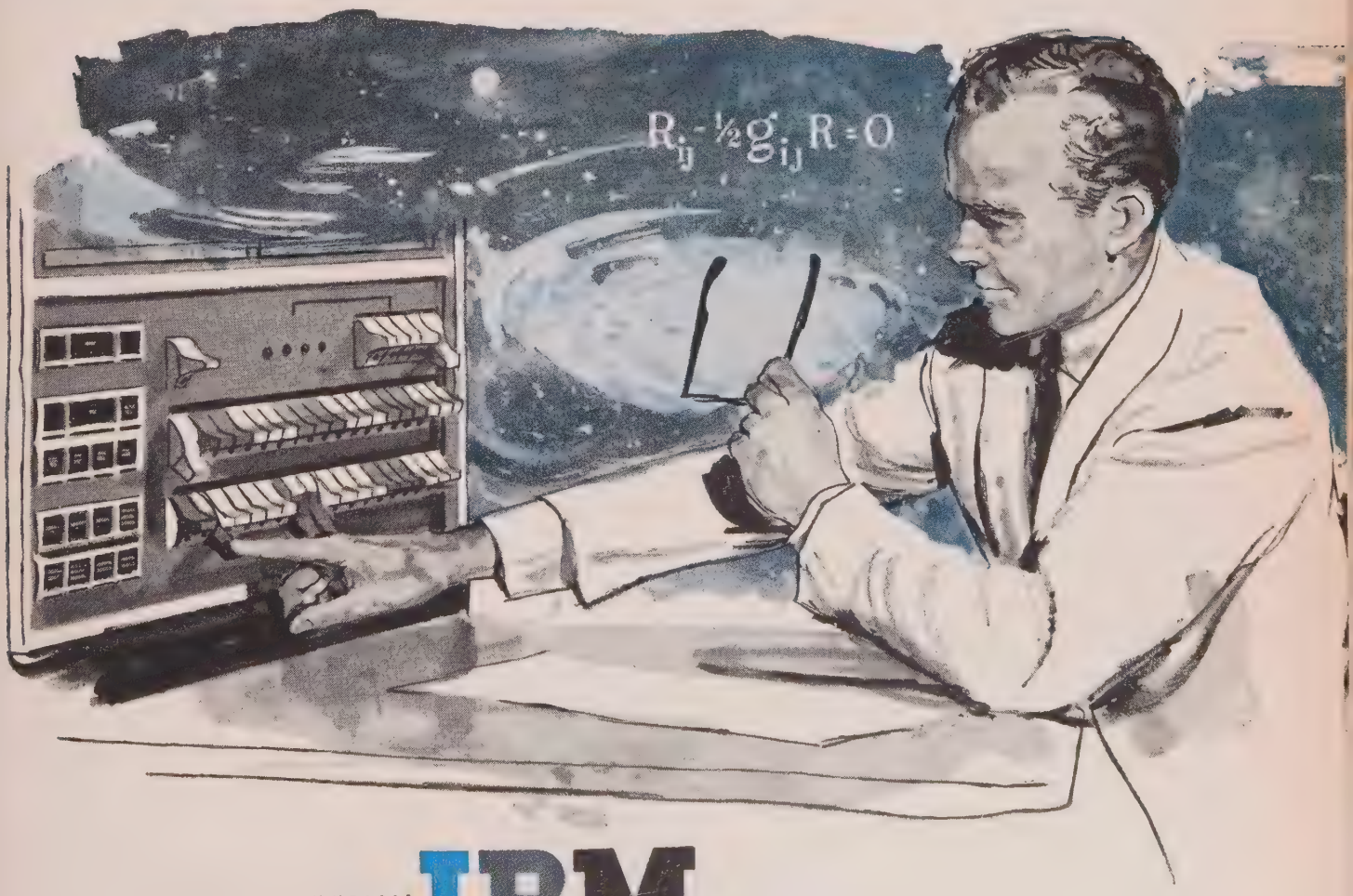

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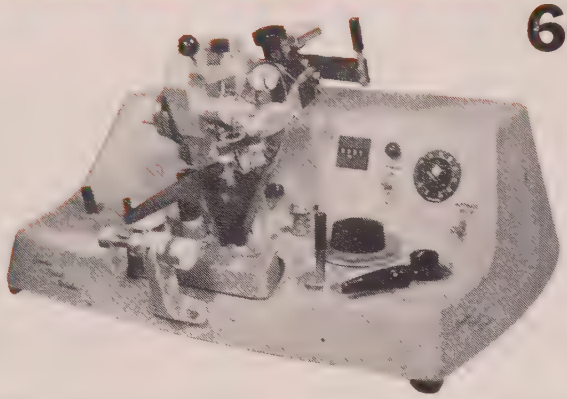

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Professional Group Meetings

(Continued from page 98A)

Electronic Computers

San Francisco—June 23

"Sequential Machines: An Introduction," Dr. D. D. Aufenkamp, General Electric Computer Lab., Palo Alto, Calif.

Industrial Electronics

Chicago—December 11, 1959

"The Infrared Radiometric Method and its Application to Remote Temperature Measurement," H. Berman, Comptometer Corp., Niles, Ill.

Chicago—May 13

"Thermoelectric Cooling, a New Field for the Electronic Engineer," G. Boesen, Borg-Warner Corp., Des Plaines, Ill.

Instrumentation

Washington, D. C.—March 14

"Recent Data from Biological & Environmental Missile-Qualified Instrumentation," R. Adams, Aerospace Medical Center, Randolph Field, Texas.

Washington, D. C.—April 4

"Project TRANSIT, a Navigation Satellite," Appl. Physics Lab., Johns Hopkins University, Baltimore, Md.

Microwave Theory & Techniques

Omaha-Lincoln—April 7

1) "Electrical Engineering Graduate Program," C. M. Hyde, Univ. of Nebraska, Lincoln.
2) "Demonstration of a Coaxial Line Parametric Amplifier," L. G. Larsen, Univ. of Nebraska.
3) "Demonstration of a Resistance Tunnel Diode Oscillator," N. A. Lindsay, Univ. of Nebraska.
4) "Demonstration of a Tunnel Diode Amplifier," V. Bollesen, Univ. of Neb.

Reliability & Quality Control

Florida West Coast—August 31

"Parameters of Reliability," C. M. Ryerson, RCA, Camden, N. J.

Space Electronics and Telemetry

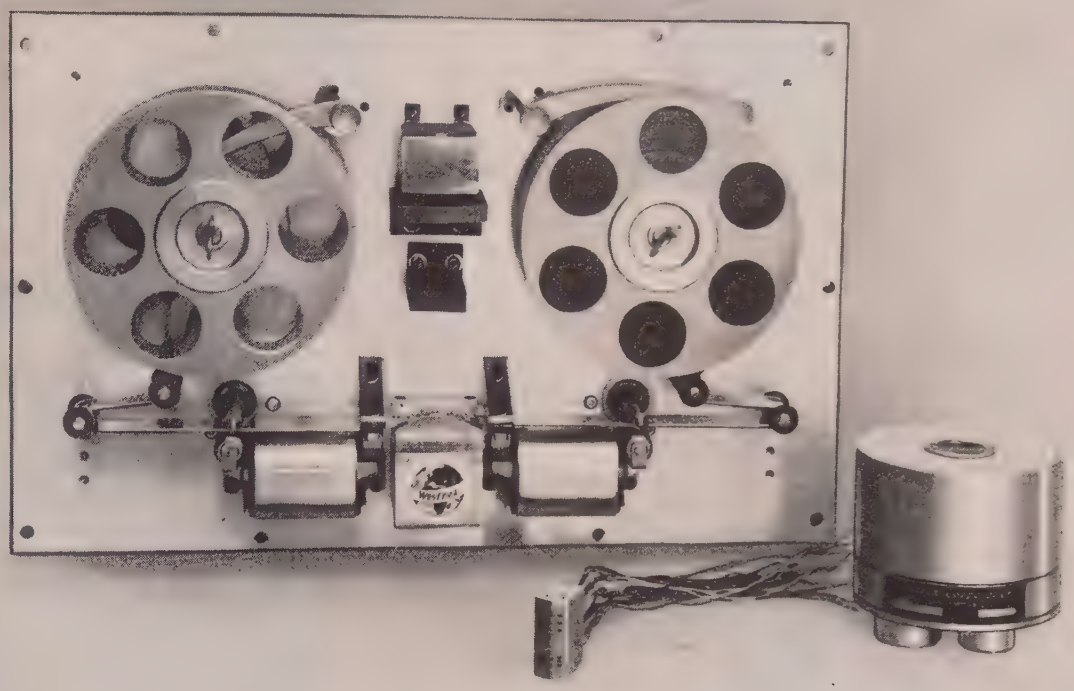
San Diego—June 23

"The Widening Horizons of Space Electronics & Telemetry," C. Hoeppner, Radiation, Inc., Melbourne, Fla.

"The Future Advance in Space Electronics," R. Werner, Cubic Corp., San Diego, Calif.

(Continued on page 106A)

A WESTREX BRIEFING... To your specifications for recording, storing, and recovering data, Westrex brings more than a quarter of a century of experience. Our major disciplines are (1) electronics, (2) mechanics, as needed in mechanical design for precise tape-pulling mechanisms, and (3) optics. Here, briefed, are descriptions of some of our new products...



PERFORATED TAPE READERS A new Westrex perforated tape reader handles 8 levels of information; reads in both directions at continuous speeds up to 1400 characters per second; stops on character at 200 characters per second. Starting time 22 milliseconds, stopping time 1 millisecond. Remote control operation and logic level output to meet your specific needs. Solid state electronics, with miniaturized etched circuit, plug-in modules. End of tape sensing by logic read out.

MAGNETIC HEADS These include multiple section instrumentation heads; memory drum heads; and erase-record-reproduce assemblies for applications that range from sound systems to missiles. Catalog items or custom-built units to your requirements. Our experience assures proper utilization of design factors that most economically meet your needs. Consideration of your special requirements, such as high crosstalk rejection, stability under extreme environmental conditions, and precise mechanical tolerances, are a part of our service to customers. What are your needs?

MINIATURE AIRBORNE TAPE RECORDERS Designed to withstand impacts of 1500 G's, a new Westrex miniature recorder can simultaneously record and monitor 14 tracks of information. With 14 tracks to the inch, unique shielding provides a crosstalk ratio of over 40 db at 5000 c.p.s. Precise gap alignment, obtained by optical lapping methods, maintains gap scatter within plus or minus 50 microinches. The positively-driven tape-pulling mechanism, and virtually continuously supported tape, are features which reflect our unique and proprietary knowledge in this field. The entire hermetically-sealed recording unit is contained in a single cylinder 3 inches high and 4 inches in diameter.

For information on this and other Westrex products, address your inquiry to Mr. L. A. Call, Westrex Corporation, Recording Department, 6601 Romaine Street, Hollywood 28, California.

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MINIATURE AUDIO TRANSFORMERS
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- Frequency response:
  ± 2DB 30-20,000 CPS
  *± 2DB 200-10,000 CPS

Cat. No.	Imped. level —	Appl.	MIL Type
PMA 1	Pri. 50/200/500 Sec. 60,000 C.T.	Line or mike to single or P.P. grids	TF4RX10YY
PMA 2	Pri. 4/8 Sec. 60,000 C.T.	Dynamic mike or spkr. voice coil to single or P.P. grids	TF4RX10YY
PMA 3	Pri. 50/200/500 Sec. 60,000 C.T.	Line or mike to single or P.P. grids	TF4RX10YY
PMA 4	Pri. 15,000 Sec. 60,000 C.T.	Single triode plate to single or P.P. grids	TF4RX15YY
PMA 5*	Pri. 15,000 Sec. 60,000 C.T.	Single triode plate to P.P. grids	TF4RX15YY
PMA 6	Pri. 15,000 Sec. 50/200/500	Single triode plate to multiple line	TF4RX13YY
PMA 7*	Pri. 15,000 Sec. 50/200/500	Single triode plate to multiple line	TF4RX13YY
PMA 8	Pri 30,000 C.T. Sec. 50/200/500	Push-pull triode plate to multiple line	TF4RX13YY
PMA 9	Pri. 60,000 C.T. Sec. 50/200/500	Crystal mike or pickup to multiple line	TF4RX13YY
PMA 10	Pri. 50/200 Sec. 50/200/500	Mixing or matching	TF4RX16YY
PMA 11	40 hy, 3 ma. d.c. 3500 Ω d.c. res.	Parallel feed reactor	TF4RX20YY

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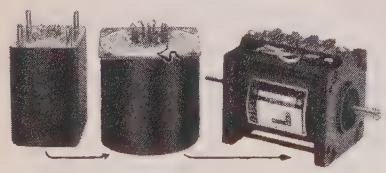

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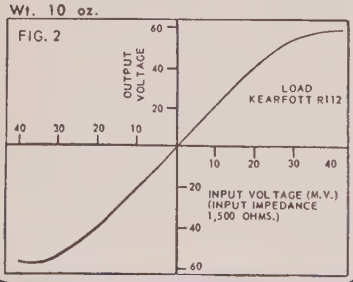

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Professional Group Meetings

(Continued from page 104A)

VEHICULAR COMMUNICATIONS

Chicago—September 9

"Personal Radio Pag ng in the VHF Band," J. Mitchell, Motorola Inc., Chicago, Ill.

Section Meetings

ANCHORAGE

Farewell Dinner for Messrs. R. O. Voight and J. M. Walden. 8/29/60.

CLEVELAND

"Economic & Political Implications of Electronic Development Abroad," S. J. Begun, Clevite Corp.; Executive Committee Meeting. 9/8/60.

DALLAS

"Design & Evaluation of Microwave Anechoic Chambers," E. F. Buckley, Emerson & Cuming, Inc. 9/20/60.

DETROIT

"Electronics in Astrobiology," T. C. Helvey, Ortronix, Inc. 9/16/60.

ERIE

"The Mechanical Cricket," Phil Sterner, Erie Resistor Corp.; "An Integrating Thermometer," Allen Davidson, Allen Davidson Elec. Development Co.; "Mathematical Approaches to New Designs of Old Components" & "Method of Equivalent Linearization of Kryloff & Bogoliuboff," Bob Gray. 5/19/60. "The IRE Member's Duties & Responsibilities," J. N. Dyer, Airborne Instruments Lab. 6/15/60.

EVANSVILLE-OWENSBORO

"Electronic Cooking," G. W. Schroeder, GE Co. 9/14/60.

FLORIDA WEST COAST

"Aspects of Stable Platforms & Inertial Guidance as Applied in Earth & Space Navigation," M. P. Smith, Minneapolis-Honeywell Regulator Co.; "Basic Principles of Ferrite Devices as Applied to Microwave Techniques," John Oliver, Sperry Microwave Electronics Co. 9/21/60.

HAWAII

"Radio Communications With Submarines," H. A. Wheeler, Wheeler Labs. 4/13/60.

"Radio Astronomy," Grote Reber, Research Corp. 5/11/60.

Annual Social Meeting—6/10/60.

"Discoverer Program & The Tracking of Satellites," W. E. Mileski, Lockheed Corp. Joint meeting with American Rocket Society. 8/10/60.

KANSAS CITY

Annual Picnic. 6/25/60.

LITTLE ROCK

"Your Institute of Radio Engineers," C. E. Harp, University of Oklahoma; Film: "Easy Street." 9/12/60.

(Continued on page 108A)

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TECHNICAL BOOKS

AIRBORNE RADAR SYSTEMS
by Donald J. Povejsil, Robert S. Raven and Peter Waterman

Provides an understanding of basic radar technology and its relation to overall weapons system design, with emphasis on basic principles and systems analysis techniques and how mathematical models may be developed to solve radar design problems. *Principles of Guided Missile Design series.*
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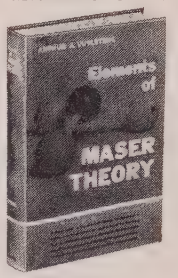

A basic guide to the physical theories fundamental of maser theory including all of the concepts and mathematical apparatus necessary to an understanding of the basic subject matter.
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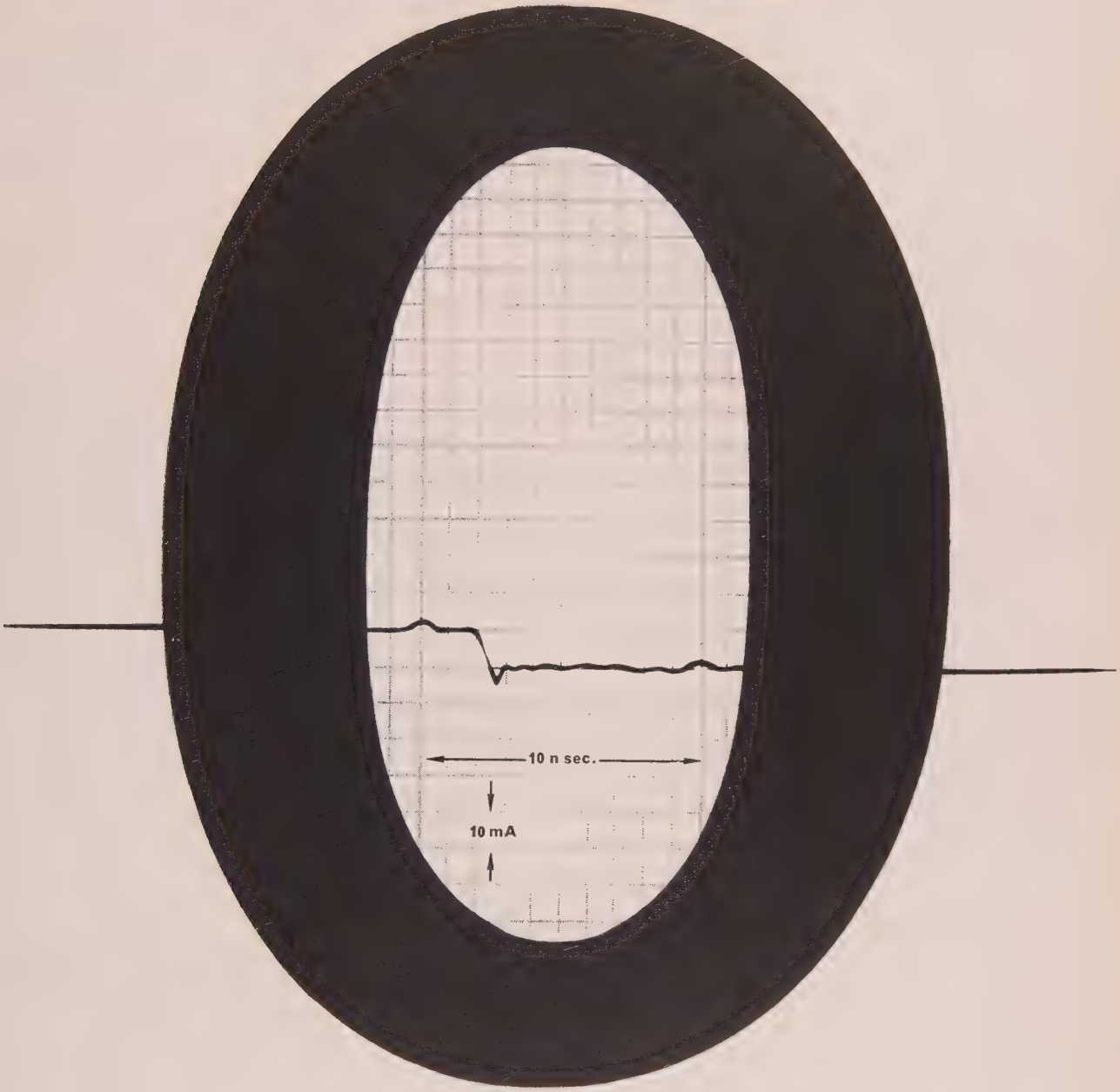

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Type	I_F@1V (mA)	E_S@100μA (volts)	I_R(μA) @ −5V @25°C	@100°C	Recovery n sec.
HD 5000	5	20	0.2	2.0	<0.5
HD 5001	5	20	1.0	10.0	<0.5
HD 5002	2	20	0.2	2.0	<0.5
HD 5003	2	20	1.0	10.0	<0.5
HD 5004	2	15	1.0	20.0	<0.5

The recovery circuit uses a high-speed sampling scope and attachments. The switching is 10 mA forward to 6V reverse; recover to 1mA reverse. Loop impedance is 100 ohms. Typical capacitance: 0.8 pf. Typical rectification efficiency: 60% at 100 Mc.

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HUGHES AIRCRAFT COMPANY

"I owe my success to my trusty Sigma Type 22 RJC200G SIL relay"

— PORTHOS P. GIZZARD

With a sizable amount of our business due to saber-rattling on a national scale, it's heartening to discover some relay prospects among those who practice skewering each other just for fun. One of our reps recently wrote in, calling our attention to a device in which a buzzer sounds when a proper forward-moving fencing "hit" is scored. The buzzer circuit is closed by the contacts of a Sigma "22" relay, which in turn is wired to a battery and a plunger switch at the tip of the foil or épée. The inventor's name is L.A. Wortman, and he holds no lesser rank than chairman of the Electrical Weapons Committee of the Amateur Fencers League of America, as well as American Delegate to the Electrical Signaling Comm., Federation International D'Escrime.

We sincerely hope, however, that Mr. Wortman shows more mercy in a *salle d' armes* than he does to the hermetically sealed enclosure of his Sigma sensitive relay. In describing his ingenious boon to practice fencers (fencing practicers?), he calmly states "The relay is a Sigma Type 22J200 or equivalent... (These dual series) coils must be separated and reconnected ...The case of the relay is easily removed with a pair of diagonal wire cutters. Starting at the bottom edge and peeling, the cover comes off as though it were a sardine-can cover." *Really*, Mr. Wortman. If Series 22 relay enclosures were meant to be removable, we would have made them that way. (On second thought, maybe supplying a little key with each hermetically sealed Sigma relay might not be such a bad idea at that. Remember that Air Force captain and his little drill?)

At all events, this clearly points out one fact: clever people are still successfully applying Sigma relays in ways which turn our application engineers green (92 parts horror, 8 parts envy). We can only hope that future builders of electrical fencing instruments and kindred souls will first ask us if we have what they want, before picking up the side-cutting pliers. It might pleasantly surprise some to see the assortment of open and sealed, single- or dual-coil, magnetic latching, big and little relays we can offer. We might even have one for Mr. Wortman's august body which would signal a hit not by a buzz on a buzzer, but simply by saying "ouch."

"22" Bulletin on request; application engineering by letter and over the phone.

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Section Meetings

(Continued from page 106A)

LONG ISLAND

"Doppler Techniques for Ground Velocity Measurements," H. C. Knauss, Raytheon. 9/13/60.

LOS ANGELES SECTION

"Non-Military Uses of Space Satellites in Communications," L. C. Tillotson, Bell Telephone Labs.; "A 24-Hour Satellite for Military Communications Relay," George Mueller, Space Technology Labs., Inc. 9/13/60.

LUBBOCK

Films: "The Big Bounce"—"The Sandis Story," L. A. Pease, Southwestern Bell Telephone Co.; Election of Officers. 9/20/60.

NORTHERN ALBERTA

Films: "The Physics of Semi-conductors" & "How the Blood Circulates." 9/13/60.

OMAHA-LINCOLN

Tour of Omaha University TV Studio & Physics Lab. Host J. G. McMillan, Omaha Univ. 4/7/60.

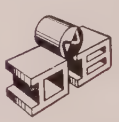
PRINCETON

Tour of Princeton-Penn Accelerator. 5/12/60.

SACRAMENTO

"Cryogenic Flow Measurements," G. R. Deppe, Aerojet-General Corp. 9/8/60.

(Continued on page 110A)



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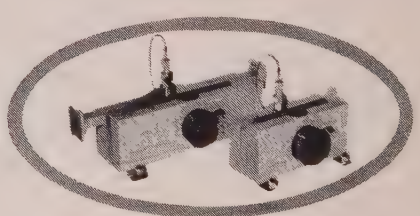

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UNPRECEDENTED SWITCH UNIFORMITY FROM ENTIRELY NEW AUTOMATED MANUFACTURING CONCEPT

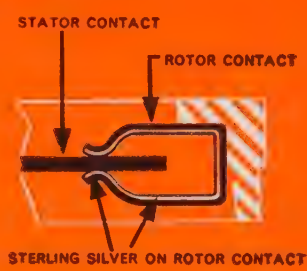
- Drift virtually eliminated by molding terminals into exact, immovable, permanent position in stator, entirely by machine . . . making all switches identical in each production run. Machine exactitude replaces human error and variations, eliminating the wider tolerances characteristic of hand assembly.
- Delicate switch parts are not exposed to breakage; many parts are not even handled.
- Superior insulation due to repetitive exactness in terminal spacing and molded glass alkyd material.
- Heat from soldering cannot loosen terminals.
- Natural design barrier prevents solder from flowing into circuit elements during soldering.

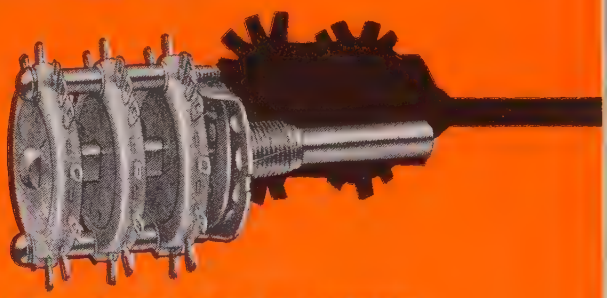
SAFER, QUICKER, EASIER HANDLING IN YOUR PRODUCTION LINE

Molded glass alkyd stator won't break or crush during ordinary handling or if accidentally dropped. Terminals cannot be dislodged or moved. Both stationary and movable contacts are enclosed for additional safety and handling convenience.

LONGER LIFE CONTACTS

- After 1,200,000 index operations, life tests reveal virtually no change in rotor contact resistance.
- 1000% *more* silver than used by most commercial switches has been *evenly* applied at actual point of contact by exclusive process. (See Sketch). Optional solid sterling silver contacts and stator available for additional contact performance.
- Rotor contacts are double-wiping and self-cleaning.
- Stator contacts are integral with terminals.





NOW! NEWLY TOOLED CLUSTER AND 3-FINGER CONTACTS

Offered for numerous additional circuits not previously available. Tremendous range of options meet most requirements without special tooling.

OTHER FEATURES

- Series 212 is available in numerous combinations with CTS variable resistors and power snap switches.
- Condensed assembly. Wafers can be stacked adjacent to each other. No spacers required.
- Non-toxic combustion fumes—an important advantage in submarine or similar closed space applications.
- Exclusive balanced detent mechanism has 2 dissimilar detent forms for definite feel and long pull in.
- New balanced lever arm and star wheel detent assembly free from end and side thrust for extremely long life and accurately controlled torque. Optional at extra cost.

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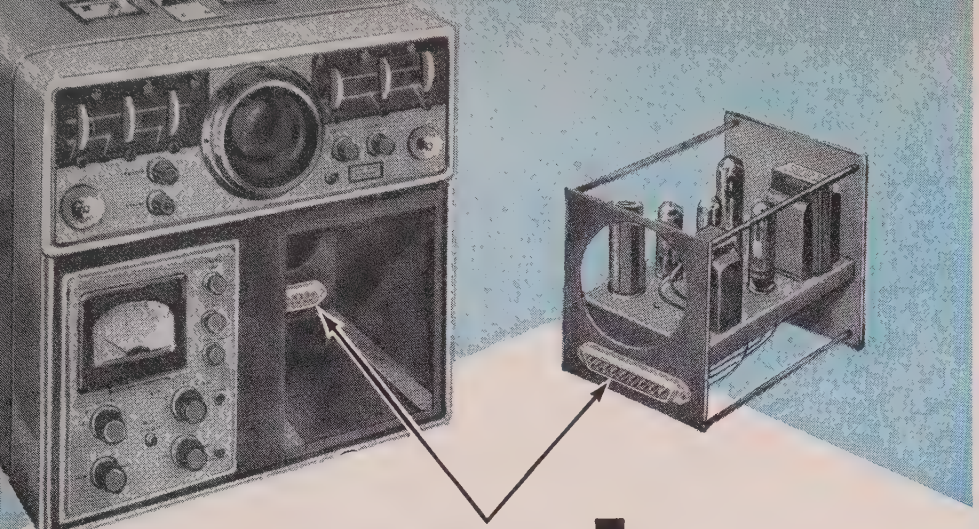
CTS *Corporation*

5 FACTORIES TO SERVE YOU: CTS Corporation, Elkhart, Indiana; Chicago Telephone of Calif., Inc., So. Pasadena, California; CTS of Asheville, Inc., Skyland, No. Carolina; CTS of Berne, Berne, Indiana; C. C. Meredith & Co., Ltd., Streetsville, Ontario, Canada. Sales



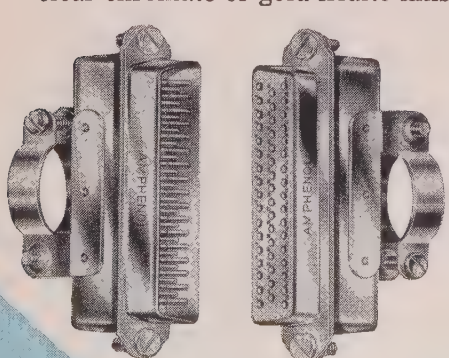
new!

Min Rac 17

miniature rack & panel connectors with POKE HOME® contacts

Solve space, weight and size problems with AMPHENOL's new Min Rac 17 connectors, true miniatures with the "Big Plus" advantage of Poke Home contacts! Min Rac 17's are rack & panel connectors ideally suited for today's compact chassis designs, connectors half the size and weight of standards, delivering full size efficiency. And with the patented Poke Home contact concept (U.S. Pat. 2,419,018), Min Rac 17's are easily, reliably assembled—contacts are crimped or soldered outside the connector body, then "poked home" for assembly.

Min Rac 17's are available in 9, 15, 25, 37 and 50 contacts in rack & panel, cable-to-chassis and cable-to-cable designs. Contacts are gold plated. Shells may be ordered with clear chromate or gold iridite finish.



These remarkable connectors are available now—write for full catalog!

AMPHENOL CONNECTOR DIVISION

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Amphenol-Borg Electronics Corporation

Section Meetings

(Continued from page 108A)

SAN DIEGO

"Research at General Atomic in the Nuclear Age," P. H. Miller, Jr., General Atomic. 8/3/60.

TOLEDO

Tour of New WSPD-TV Transmitter. 5/19/60.

Movies by Ohio Bell, "Nike Missile Defense System," "The Bell Solar Battery"; Election of Officers. 6/9/60.

WICHITA

"Inertial Navigation," Jarrett Thomas, Boeing Airplane Co. 3/23/60.

"Infrared Principles & Technology." W. J. Beyen, Texas Instruments; Joint meeting with AIEE. 4/21/60.

Presentation of Officers. 6/29/60.

"The Coming Applications of the Infidelity of HiFi (?) & Stereophonics to Missile & Aircraft Guidance Systems," C. R. Bonnell, Beech Aircraft Co.; Joint meeting with AIEE. 5/25/60.

SUBSECTIONS

NORTHERN VERMONT

"Radiological Physics in Medicine," C. D. Janney, Vermont College of Medicine. 9/19/60.

Membership

The following transfers and admissions have been approved and are now effective:

Transfer to Senior Member

Affrunti, A. J., Morton Grove, Ill.
Bacon, E. J., Rockville, Md.
Barber, L. E., Royal Oak, Mich.
Bauer, W. M., Monterey, Calif.
Bellville, W. H., Cedar Rapids, Iowa
Bickel, R. C., Homewood, Ill.
Calderhead, H. J., Euclid, Ohio
Christoffers, W. H., Los Angeles, Calif.
Collin, R. E., Cleveland, Ohio
Cooper, G., Chestnut Hill, Mass.
Culnan, D. E., San Diego, Calif.
Dax, P. R., Baltimore, Md.
De Camp, P. W., Fairborn, Ohio
Demaree, F. E., Chicago, Ill.
Doherty, J. E., Jr., Dallas, Tex.
Durkee, C. E., Highland Park, Ill.
Emery, R. H., Falls Church, Va.
Fulton, W. L., Dayton, Ohio
Fuzesi, Z., Rio de Janeiro, Brasil
Grossman, H. W., Emerson, N. J.
Hardin, C. D., Rockville, Md.
Hauser, J. H., San Diego, Calif.
Havel, J. M., Little Silver, N. J.
Hines, M. E., Burlington, Mass.
Horgan, J. D., Milwaukee, Wis.
Howard, L. L., Dallas, Tex.
Izzo, A. J., Los Angeles, Calif.
Kaelin, B. A., North Massapequa, L. I., N. Y.
Katz, L., Eatontown, N. J.
King, T. J., Sunnyvale, Calif.
Laughlin, J. G., Whittier, Calif.
Leftwich, L. L., Trabuco Canyon, Calif.
Looney, C. H., Jr., Washington, D. C.
Madey, R., Bellport, N. Y.
Mason, C. F., Fort Wayne, Ind.
Nelson, J. L., Highland Park, N. J.
Pear, C. B., Jr., Silver Spring, Md.

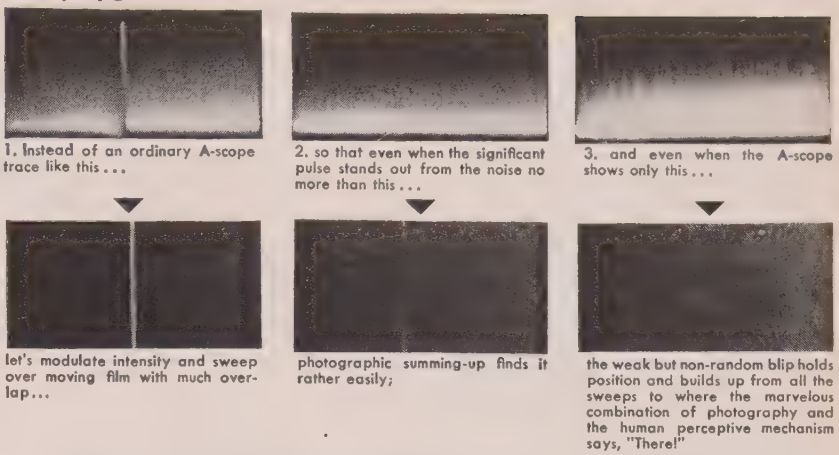
(Continued on page 112A)

Kodak reports on:

seeing the signal among the noise . . . materials that are not blinded by hot wind . . . polyester electrically upgraded

A human talent

The June issue of *Scientific American* contained an article full of learned speculation on the neurological mechanism by which lines, straight and curved, are perceived. Whatever the mechanism, the fact remains that the nervous system is very good at seeing a line from exceedingly faint physical stimuli. We had been thinking about this a lot and about ways it could help solve the nasty signal-to-noise problem that keeps cropping up on such occasions as when defense from submarine attack is considered. Today's almost instantly available photographic images make a fine bridge from an electronic system to a human nervous system. For example:



1. Instead of an ordinary A-scope trace like this . . .

2. so that even when the significant pulse stands out from the noise no more than this . . .

3. and even when the A-scope shows only this . . .

let's modulate intensity and sweep over moving film with much overlap . . .

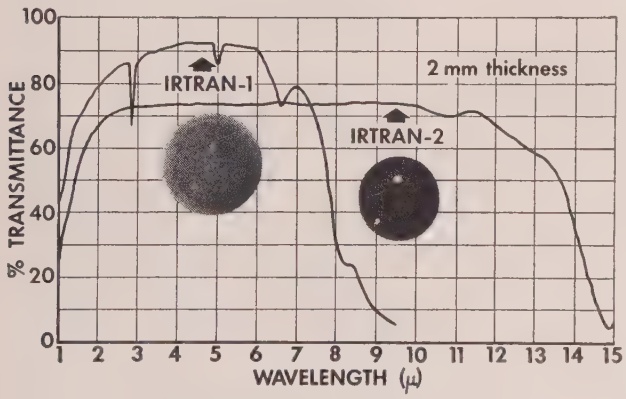
photographic summing-up finds it rather easily;

the weak but non-random blip holds position and builds up from all the sweeps to where the marvelous combination of photography and the human perceptive mechanism says, "There!"

Organizations active in military developments who wish to know more about this work should communicate with Eastman Kodak Company, Apparatus and Optical Division, Rochester 4, N. Y.

For little pots or big peepholes

If you want to make a useful little pot for putting things in that are to be poked with infrared or microwave radiation; or if you want to make an object move through the air at speeds like Mach 6 and need something to keep the hot wind out of the peephole while admitting infrared *and* microwaves to steer by, you must investigate *Kodak Irtran* material or face charges of that heinous sin, ignoring the "state of the art."



Kodak Irtran merchandise: an *Irtran-1* dome and an f/1.0 *Irtran-2* molded asphere

There are now two kinds of *Irtran* optical material, neither of them intended for wavelengths that your old dictionary would call "optical."

The newer, *Irtran-2*, has as its point of pride the transmittance curve displayed above, combined with a remarkable mechanical and chemical ruggedness. Its refractive index is around 2.2. Note the tremendous wavelength span over which transmittance losses are nearly all due to reflection. Heating to 600°C in air does nothing more than improve the transmittance—by formation of an anti-reflection coating. We can apply a much better coating, though, by evaporation.

*Irtran-1** material, the other one, needs no anti-reflection coating because its refractive index is only 1.38 at 1μ. *Its* big glory, aside from high infrared transmittance at 2 to 7μ even when very hot, stems from a 9.4 kmc dielectric constant of 5 and a loss tangent of 10^{-4}.

Decibel loss per meter for low-loss materials is

$$\frac{8.686\pi}{\lambda_o}\sqrt{K}\tan\delta$$

where K = dielectric constant
λ_o = wavelength in vacuum
tan δ = loss tangent
δ is the complement of the vector angle whose cosine the power engineer calls "power factor."

One untuned *Irtran-1* sample .012" thick we tested in the X-band frequency range introduced an attenuation of less than 0.3db and exhibited a maximum standing wave ratio of 1.5 over the band.

The ideal radome material would have a kilomegacycle dielectric constant of unity, i.e. to microwaves it would be the same as nothing at all. Plastics commonly used for radomes run around 2 or 3, but they don't transmit infrared rays. On the other hand, there are materials that are good for infrared transmission but have dielectric constant around 13.

Another way the *Irtran* materials shine figuratively is their very low tendency to glow literally when hot and thus to blind themselves by swamping transmittance with emissivity. They survive undamaged through the conventional torture tests of temperature-humidity cycling for 28 days, vacuum-steam pressure for 2,000 cycles in saline solution, sand and dust blast, acids, alkalis, organic solvents, plunges from hot to cold, etc.

Currently we can supply either of these polycrystalline Irtran *materials as optical elements up to 6½" in diameter, among other forms. Let us have no jurisdictional disputes among optickers, micro-wavers, and infraredders as to who sends for the supporting data from Eastman Kodak Company, Special Products Division, Rochester 4, N. Y.*

New recipe

We have a new polyester resin recipe. It results in a dielectric constant that is really constant up to 130° C, which could not be said of earlier polyester. Its dissipation factor is 0.3% at 100° C, where the old ran to 1.2%. Since it also better resists humidity, acid, and bases, it is excellent for insulation. As a capacitor film at 2000 v/mil, it outlasts the old 8 to 1.

Eastman Chemical Products, Inc., our subsidiary, sells the resin. Acme Backing Corporation, Canal and Ludlow Streets, Stamford, Conn., turns it to what others call film and we (for whom "film" has another meaning) call sheeting. Acme will gladly expatiate.

Kodak
TRADEMARK

*Formerly designated "Irtran AB-1."

This is another advertisement where Eastman Kodak Company probes at random for mutual interests and occasionally a little revenue from those whose work has something to do with science

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(Continued from page 110A)

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Power, R. L., Littlerock, Calif.
Ramey, R. L., Charlottesville, Va.
Schneider, R. P., King of Prussia, Pa.
Seaward, P. E., Palo Alto, Calif.
Shea, G. J., Elmhurst, Ill.
Stuhlbarg, S. M., Indianapolis, Ind.
Taylor, W. G., San Diego, Calif.
Thomas, J. B., Princeton, N. J.
Unz, H., Lawrence, Kan.
Voight, R. O., Arlington, Va.
Wenger, F. E., Washington, D. C.
Wolf, A. A., Rochester, N. Y.

Admission to Senior Member

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Baum, C. C., Santa Barbara, Calif.
Boaz, R. O., Colorado Springs, Colo.
Bolman, R. G., Canoga Park, Calif.
Brean, J. W., Cincinnati, Ohio
Bridges, A. P., Colorado Springs, Colo.
Coville, H. A., Fairview Village, Pa.
Crabbe, F. J., San Francisco, Calif.
Cushing, V. J., Rockville, Md.
Davies, D. L., Dorking, Surrey, England
Del Toro, V., New York, N. Y.
Ellis, H. K., Vestal, N. Y.
Feick, W., Valparaiso, Chile
Gootee, T. E., Falls Church, Va.
Gore, W. L., Newark, Del.
Gruber, A. L., El Segundo, Calif.
Guggenbuhl, W., Zurich, Switzerland
Guier, W. H., Silver Spring, Md.

(Continued on page 114A)

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AVNET ELECTRONICS CORP.

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- ☐ Bendix Scintilla Connectors
- ☐ U. S. Semcor Semiconductors
- ☐ Sangamo Capacitors
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NOTABLE ACHIEVEMENTS AT JPL...

pioneering work in worldwide communication via satellites

PROJECT

ECHO

On August 12th, 1960, JPL scientists at Goldstone, California, radioed the world's first transcontinental microwave message to be relayed by a passive, artificial earth satellite. This satellite was the 100 ft. plastic balloon Echo I orbiting around the Earth at an altitude of approximately 1,000 miles. A pre-recorded statement by President Eisenhower was received 2,300 land miles away by scientists of the Bell Telephone Laboratories at Holmdel, New Jersey, as clearly as any conventional telephone call, in a fraction of a second.

Later in the course of the Echo experiment, the scientists at Goldstone and Holmdel conducted 2-way voice communication via the balloon satellite, Goldstone transmitting at 2,390 megacycles and Holmdel at 960 megacycles.

This successful experiment has demonstrated the feasibility of worldwide communication and is typical of many pioneering achievements of the Jet Propulsion Laboratory.

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A Research Facility operated for the National Aeronautics and Space Administration

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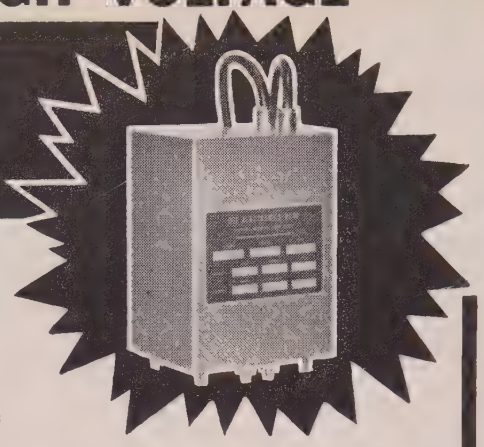
Send professional resume, with full qualifications and experience, for our immediate consideration

COMPACT HIGH VOLTAGE POWER SUPPLIES



- *Selenium Rectifiers*
- *Low Ripple*
- *Reversible Polarity*
- *Shielded Output Cables*
- *Hermetically Sealed Can*
- *Epoxy Filled*
- *INPUT: 115 Volts—60 to 400 Cycles*

Model No.	*Output*	*Dimension* W	D	H
2.5-2-1	2.5KV-2MA	3-1/16 x	2⅝	x 4¼
5-2-1	5KV-2MA	3-9/16 x	3-1/16 x	3⅞
10-1-1	10KV-1.25MA	3-9/16 x	3-1/16 x	4⅞
15-1-1	15KV-1.25MA	4	x 5	x 5¾
20-1-1	20KV-1.25MA	4¼	x 5⅛	x 6¾
25-1-1	25KV-1.25MA	4¾	x 6	x 7¼
30-1-1	30KV-1.25MA	5	x 6½	x 7¾
40-2-1	40KV-2MA	8¼	x 6¼	x 7

REGULATED SUPPLIES
Transistorized TRHV line of compact supplies from 1 KV at 5 MA to 20 KV at 1 MA available from stock.

ALL DEL products represent the ultimate in quality. Their design and manufacture incorporate only the finest materials and workmanship to assure thoroughly dependable operation in industrial and research applications.

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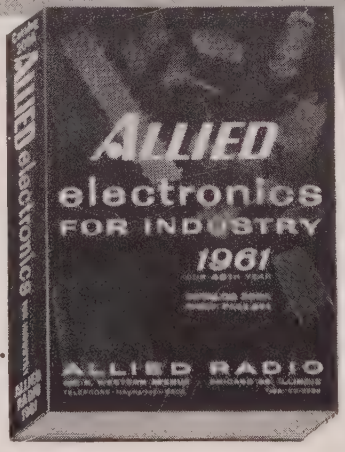

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(Continued from page 112A)

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Hormats, M. A., Patrick AFB N. M.
Ian, D. C., Pelham Manor, N. Y.
Itoh, M., New York, N. Y.
Kershner, R. H., Baltimore, Md.
Kirsten, B., Annapolis, Md.
Knight, T. G., North Redding, Mass.
Knopp, A. M., Arlington, Va.
Lewis, B. L., Winter Park, Fla.
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Litman, S., Columbia, S. C.
Litz, F. A., San Jose, Calif.
Mackiernan, D. W., Levittown, Pa.
McKenzie, J. B., Burbank, Calif.
Medler, E. H., Granada Hills, Calif.
Messerle, H. K., Sydney, N.S.W., Australia
Muroga, S., Yorktown Heights, N. Y.
Nelson, I. W., Decatur, Ill.
Norris, P. C., Morton Grove, Ill.
Paling, A. W. M., Hilversum, Netherlands
Pilgrim, K. A. W., Washington, D. C.
Putnaerglis, R., Ottawa, Ontario, Canada
Reid, D. A., Colorado Springs, Colo.
Richardson, D. K., Los Angeles, Calif.
St. John, G. E., Burlington, Mass.
Sarley, J. M., Binghamton, N. Y.
Schindler, M. J., Harrison, N. J.
Scott, C. D., Los Angeles, Calif.
Shalag, W. C., Springfield, Va.
Smith, N. B., Alden, N. Y.
Spiglanin, E., La Mirada, Calif.
Strohm, W. B., Hopewell Junction, N. Y.
Tapp, J. G., San Gabriel, Calif.
Taynton, R., Edgewater Park, N. J.
Warren, J., Irvington, N. Y.
Weill, G. G., Los Angeles, Calif.
West, H. R., Rockville, Md.
Whitcombe, D. W., Creve Coeur, Mo.
Williams, M. E. L., Djakarta, Indonesia
Wolontis, V. M., Murray Hill, N. J.
Zimmermann, R. E., New York, N. Y.

Transfer to Member

Allen, A. E., Liverpool, N. Y.
Allen, R. J., Waterloo, N. Y.
Amante, J. J., Roslindale, Mass.
Ammon, G. M., Haddonfield, N. J.
Anderson, A. F., Syracuse, N. Y.
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Berberek, W. J., Summit, Ill.
Bergeron, J. A., Nashua, N. H.
Biagiotti, R. J., Millis, Mass.
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Bunting, J. H., West Acton, Mass.
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Callaghan, E. P., New Hyde Park, L. I., N. Y.
Cardozo, D. J., North Westport, Mass.
Carlson, A. W., Harrison, Me.
Carson, R. E., Rochester, N. Y.
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Cohen, S. A., Sunnyvale, Calif.
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Collard, C. A., Garden City, L. I., N. Y.
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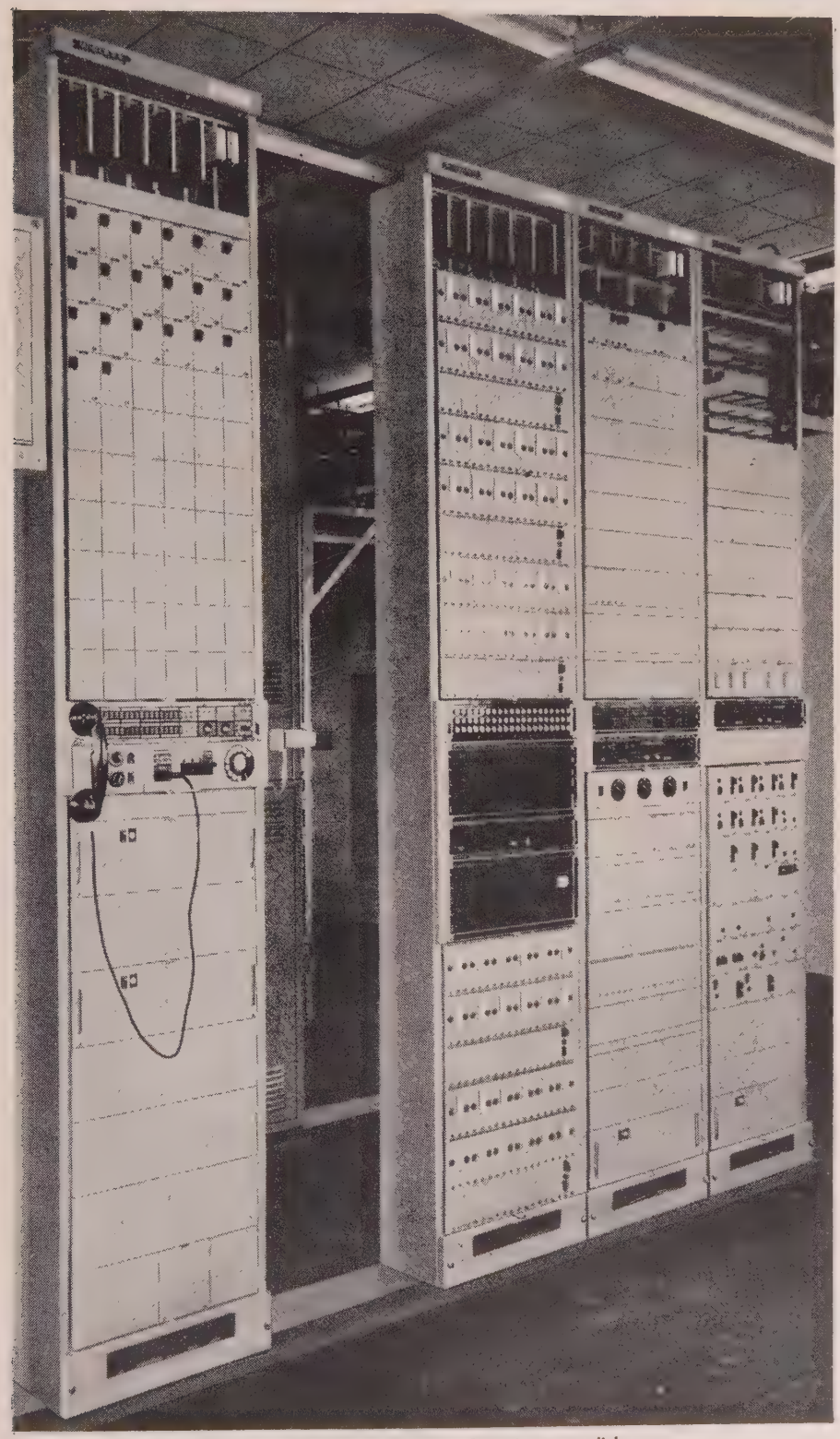
Cranshaw, R. J., Jr., Westover AFB, Mass.
Crissey, D. L., Jamestown, N. Y.
Cummings, C. S., 11, Easton, Conn.
Damron, D. H., Johnson City, N. Y.
Darois, P. F., Ridgefield, Conn.
Dawson, W. C., Fort Washington, Pa.
De Bolt, H. E., Syosset, L. I., N. Y.

(Continued on page 116A)

NEC microwave carrier terminal equipment with transistor reliability



4-bay array of NEC 120-channel terminal equipment for microwave links

This fully transistorized carrier telephone equipment for microwave links provides 120 toll quality telephone channels. Characteristics meet or surpass all CCITT requirements.

It offers these advantages over conventional tubed equipment:

* 90% reduction in power consumption
* Lower operating temperature
* Less floor space
* Increased reliability, resulting in lower maintenance costs.

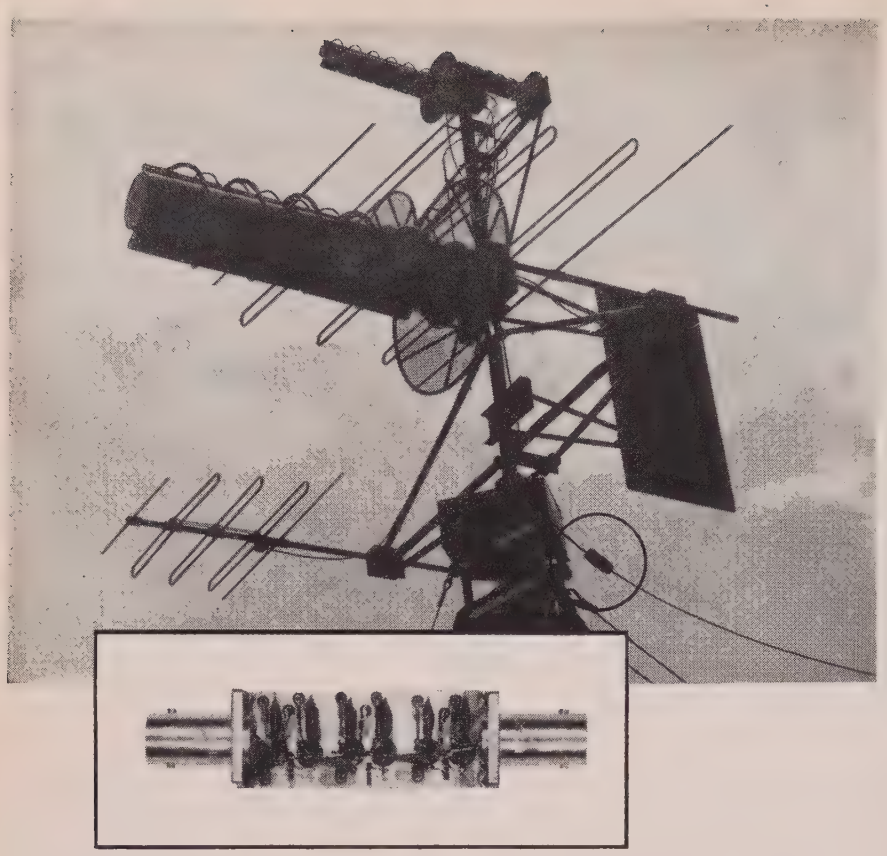
Each 300 – 3,400 c/s telephone channel can accommodate 24 voice-frequency telegraph channels operated at a keying speed of 50 bauds.

NEC will also make available within this year 600-channel transistorized equipment.

A part of print-wired channel translating unit

NEC Nippon Electric Co., Ltd. Tokyo, Japan

Communications Systems / Electronics



First Transistorized Amplifier With Bandwidth From 25 kc to 600 mc

another design problem solved by HRB-SINGER

Need for compact broadband amplifiers with increasingly wider bandwidths has resulted in HRB-SINGER's development of new circuit techniques in transistorized amplifiers.

The 25 kilocycles to 600 megacycles amplifier, no larger than half a cigarette case, covers over fifteen octaves with high gain and high efficiency. It is representative of the outstanding achievements of HRB-SINGER in this field. Reliability, ruggedness and low power consumption make these new amplifiers ideal for applications in airborne receiving equipment, portable receivers and other specialized applications.

This transistorized amplifier is only one of many outstanding amplifier models now available. In addition, HRB-SINGER has the ability and know-how to meet your special requirements, whether for a single amplifier or a complete receiving system. Direct your inquiries to Dept. G-10. A comprehensive series of data sheets describing this capability is yours for the asking.

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(Continued from page 114A)

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Deranian, M., Malden, Mass.
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(Continued on page 118A)

LAMBDA
COM-PAK Series
Regulated Power Supplies

- Immediate delivery from stock.
- Convection cooled, no troublesome blowers.
- Rated for 24 hour continuous operation at 50°C ambient.
- Swing open back for easy access.
- Transient free-output.
- Excellent regulation.

Guaranteed 5 Years

C-280 (without meters):	0-200 VDC, 0-200 MA. $184.50	C-880 (without meters):	0-200 VDC, 0-800 MA. $340.00
C-281 (without meters):	125-325 VDC, 0-200 MA. 159.50	C-881 (without meters):	125-325 VDC, 0-800 MA. 315.00
C-282 (without meters):	325-525 VDC, 0-200 MA. 169.50	C-882 (without meters):	325-525 VDC, 0-800 MA. 360.00
C-480 (without meters):	0-200 VDC, 0-400 MA. 259.50	C-1580 (without meters):	0-200 VDC, 0-1500 MA. 550.00
C-481 (without meters):	125-325 VDC, 0-400 MA. 244.50	C-1581 (without meters):	125-325 VDC, 0-1500 MA. 575.00
C-482 (without meters):	325-525 VDC, 0-400 MA. 259.50	C-1582 (without meters):	325-525 VDC, 0-1500 MA. 650.00

For Com-Pak Series models with meters, add the suffix "M" to the model number and add $30 to the unmetered model price.

COM-PAK FEATURES

- Hermetically-sealed transformer—designed to MIL-T-27A
- 50 to 400 CPS input
- Semi-conductor rectifiers for higher efficiency in C-400, C-800 and C-1500 series.
- Overload protection with built-in blown-fuse indicators
- Stable, low noise wire wound reference networks and multipliers
- Oil-filled, hermetically-sealed capacitors
- Conservatively rated for continuous duty
- Easy service access without removal from rack; all tubes readily accessible for replacement
- Provision for remote DC vernier voltage adjustment
- Exclusive design, height only 5¼″ (C-200 and C-400 series), 7″ (C-800 series) and 8¾″ (C-1500 series)

CONDENSED DATA

Regulation: *Line* Better than 0.15% or 0.3 Volt (whichever is greater). For input variations from 105-125 VAC.

Load Better than 0.25% or 0.5 Volt (whichever is greater). For load variations from 0 to full load.

Transient Response: *Line* Output voltage is constant within regulation specifications for step-function line voltage change of plus (+) 10 volts or minus (−) 10 volts rms within the limits of 105-125 VAC.

Load Output voltage is constant within regulation specifications for step-function load change from 0 to full load or full load to 0.

Ripple and Noise: Less than 3 millivolts rms.

AC Output: (unregulated) . . . 6.5 VAC (at 115 VAC Input). C-200 Series . . . 10 AMP; C-400 Series . . . 15 AMP; C-800 Series . . . 20 AMP; C-1500 Series . . . 30 AMP.

AC Input: 105-125 VAC, 50-400 CPS

Controls: DC Output Controls: Band-switches and screw-driver adjusting vernier-control, rear of chassis.

LA 111

SEND TODAY FOR COMPLETE DATA

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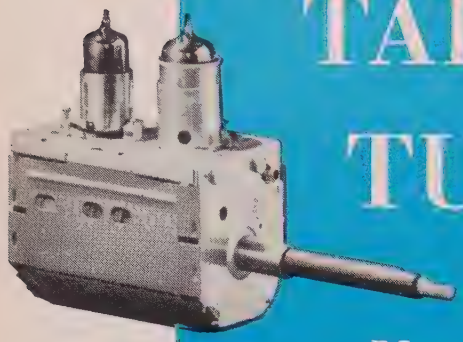
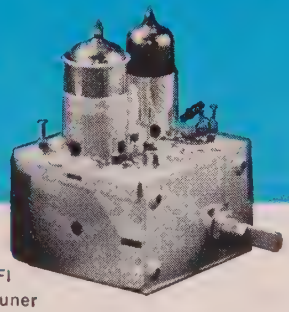
TARZIAN TUNERS



HOT ROD
(turret-type)

Noted for their DEPENDABILITY ...HIGH QUALITY and LOW COST

SILVER SEALED
(switch-type)



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(FM) Tuner

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Today TARZIAN TUNERS are backed up by more than 15 years of experience in the design . . . development . . . and production of more than 15 million TV Tuners.

And, Tarzian is the only commercial manufacturer offering both the HOT ROD (turret-type) and SILVER SEALED (switch-type) . . . as well as the newer Hi Fi FM Tuner. All embody the high standards of QUALITY . . . DEPENDABILITY . . . and PERFORMANCE that have made Tarzian a leader in the field.

For more information, write to: Sales Department
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Manufacturers of Semiconductors, Air Trimmers and Broadcast Equipment

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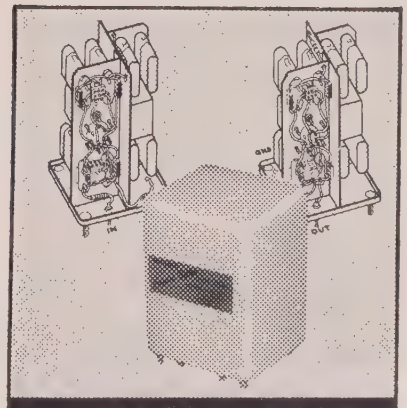
(Continued on page 120A)

BULOVA PRECISION CRYSTAL FILTERS



Whatever the frequency you wish to "isolate", Bulova experience with prototype and production quantities of precision filters assures maximum sensitivity and stability. The following examples show Bulova's mastery of the most difficult problems in high-performance filter engineering.

BAND PASS FILTERS—In a band of 30 filters, insertion loss variation between filters, and over the temperature range 25° C to 75° C, held to .3db between highest and lowest. *Part #69-A-RP-13-2N* (1 thru 30)

SINGLE SIDE BAND FILTERS—Band ripple held to ± ½db, both 1 and 3db points defined, over the temperature range 0° C to 85° C, and 300 to 2000cps vibration at 30G level. *Part #117B-FC-22-4WU*

DISCRIMINATOR—Center frequency held to within 10cps, frequencies equally spaced from center, held to 5.4v peak ± 5%. *Part #186C-TN-22A-WD*

BAND SUPPRESSION FILTERS—2kc wide band attenuated 60db, right next to it a pass band held flat to ± ¼db for 150kc. *Part #158-TF15-6R*

If you're faced with tough filtering problems, need additional information or practical application assistance, contact Bulova for engineering specialists to assist in selection of filters best suited to your needs. Write Department 1820, Bulova Electronics, Woodside 77, N.Y.

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Designers and Manufacturers of

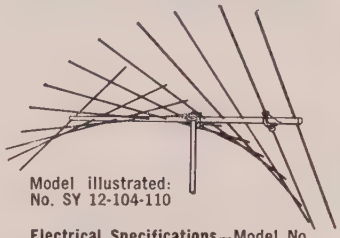
COMMERCIAL SERVICE "BEAMED-POWER" ARRAYS AND TWO-WAY SYSTEMS

Model illustrates a wide-spaced, 12 element circular polarized optimum-tuned skewed dipole "SPIRALRAY" antenna. Provides unusually high gain, even response, in all polarization planes, vertical, horizontal or oblique with unusually high signal-to-noise ratio.

NO OTHER CIRCULAR POLARIZED ARRAY known to the art today can provide the linear high gain and signal-to-noise ratio in all radiation planes.

The ideal antenna for missile tracking, telemetering and no-fade response to mobile (or moving) stations.

Models available to extend the practical range of 2-Way Communication Systems.



Model illustrated: No. SY 12-104-110

Electrical Specifications—Model No. SY-12-104-110: Polarization, circular, linear within ½ db. Gain 13 db. F/B-Ratio 30 db. V/S/W/R (50 ohm cable) 1.1/1. Beamwidth at half power points 33 degrees. Max. power input 300 w, with "Balun" supplied.

Mechanical Specifications: Boom diameter 2" O.D. x 25 ft. All aluminum boom and elements. Weight approx 25 lbs. Rated wind-load 90 mph. No ice load. Available for 120 mph wind load. (Model No. MSY-104-110).

Model SY-12-104-11 **$265.00**

Model MSY-104-110 **$390.00**

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- Telrex is equipped to design and supply to our specifications or yours, Broadband or single frequency, fixed or rotary arrays for communications, FM, TV, scatter-propagation, etc.
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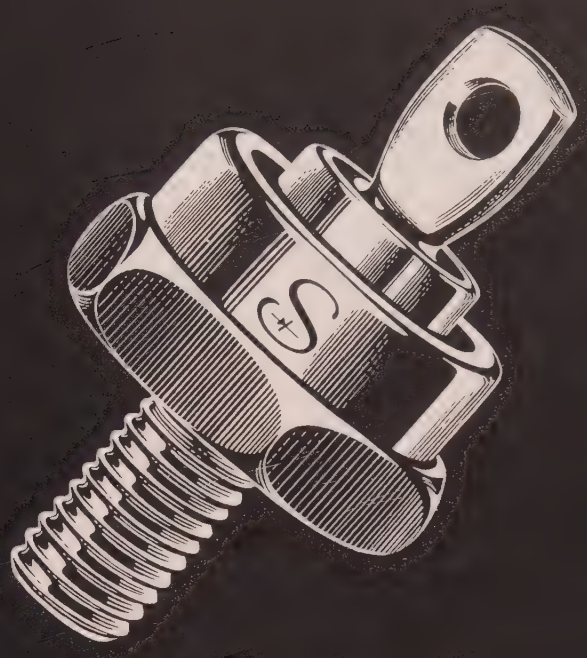
(Continued on page 122A)



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Specify SRC!

An outstanding illustration of SRC quality, reliability and safety is found in its 35 ampere silicon power rectifiers. For example, these hermetically sealed diodes are so conservatively rated that they will withstand **1200 ampere** surges!*

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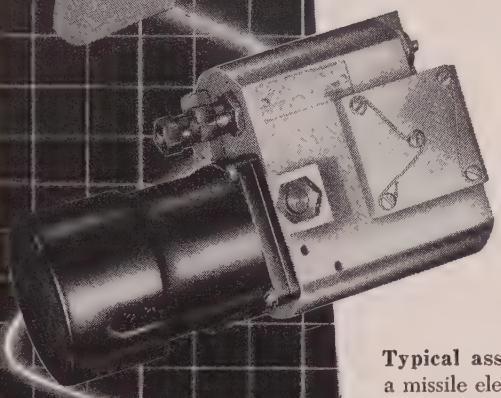

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Solution: a system designed and built around the Eastern E/HS 5 type 100A hydraulic power supply, using a 400-cycle motor. Minus motor, but including pump, reservoir, expansion chamber, pressure regulating valve and filter, the supply weighs only 2.6 lbs.

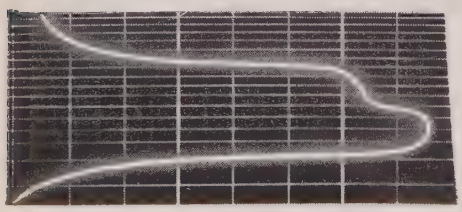
Eastern achieved a 20% saving in hydraulic power supply capacity by the use of one of its own *pulse length modulated* single-stage servo valves, manifolded to each actuator. Zero first stage leakage, zero hysteresis, plus the high response and efficiency which characterize this valve contributed to the successful meeting of the demanding requirements.

Specifications:

force level: 300 lbs.
total stroke: 1/2 in.
max. velocity: 1.5 in./sec.
time constant: 0.1 sec.
max. actuator width: 1.25 in.

For genuine contributions to the solution of your servo problems, call in the Eastern engineer.

servo-valves • fuel controls • actuators



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Kriss, G. H., Hamilton Square, N. J.
Kutelmach, M., Glenolden, Pa.
Lagoe, J. A., North Hollywood, Calif.
Lanzano, G., Washington, D. C.
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Lash, W., Los Angeles, Calif.
Lavallee, L. R., Boylston, Mass.
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Lepp, K. E., Vanderberg AFB, Calif.
Lerman, G. S., Clayton, Mo.
Lesley, A. M., Woodland Hills, Calif.
Levers, P. D., Carleton Place, Ont., Canada
LeVier, P. W., Pasadena, Calif.
Levy, N., Tel-Aviv, Israel
Lewis, C. J., Englewood, N. J.
Lovallo, F. J., Jr., Red Hook, N. Y.
Luck, T. P., Oxnard, Calif.
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Lyons, S., Burbank, Calif.
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(Continued on page 124A)

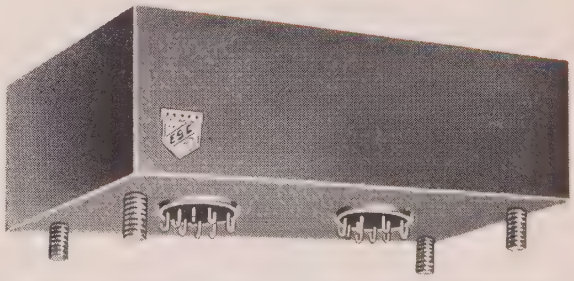
TIGHT SQUEEZE!

ESC's new miniature Transponder Delay Line fits into just 6 cubic inches!

For modern airborne equipment, where space and weight are critical, ESC has created a new **Miniature Transponder Delay Line** —Model 52-44...which embodies the most advanced techniques of weight and space reduction. It measures just 6 cubic inches total!

Specifications—Model 52-44, Lumped Constant Delay Line:

Impedance—470 ohms
Delay Time—20.3 ±.1
Rise Time—.6 (max.)
Temperature Coefficient—
65 ppm or better over a temperature range of −55°C to +125°C
Attenuation—4 db
Size—1" x 2" x 3"
Weight—6 ounces
Tapped as required



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exceptional employment opportunities for engineers experienced in computer components...excellent profit-sharing plan.

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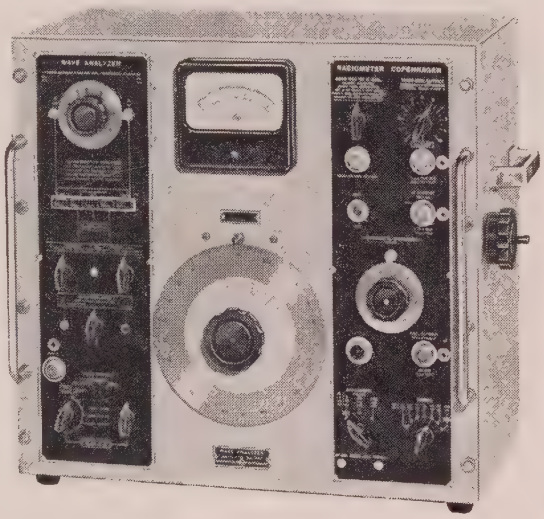
Distributed constant delay lines • Lumped-constant delay lines • Variable delay networks • Continuously variable delay lines • Step variable delay lines • Shift registers • Video transformers • Filters of all types • Pulse-forming networks • Miniature plug-in encapsulated circuit assemblies

Looking for an audio-frequency Wave Analyzer with variable bandwidth?

WAVE ANALYZER

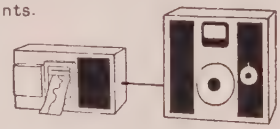
Type FRA2T



3 Bandwidths: 2 - 25 - 125 cps (÷1 db) provide easy tuning over the entire frequency range and permit measurements to be made on frequency-modulated signals or in the presence of wow.

Built-in Tone Generator (B.F.O.) for selectivity measurements with provision for external level control. The B.F.O. can be offset from the Analyzer tuning frequency by 0 to 16 kc for intermodulation measurements.

Frequency Response Recorder



The Analyzer can be connected and mechanically coupled to the Recorder type NS 3 which uses paper charts 20 cm long graduated directly in frequency and amplitude. Complete recordings in 130 secs.

Frequency Dial 0 to 16 kc with linear calibration up to 100 cps for high resolution at low frequencies, logarithmic from 100 cps to 2 kc, and linear from 2 kc to 16 kc. Accuracy 0.5% + 1 cps.

Incremental Dial from 0 to ± 60 cps, direct reading.

15 Ranges from 100 microvolts to 1000 volts f.s.: Hum and noise less than 2 microvolts, lowest detectable signal 3 microvolts. Input impedance 2 megohms on all ranges.

Write for complete information.

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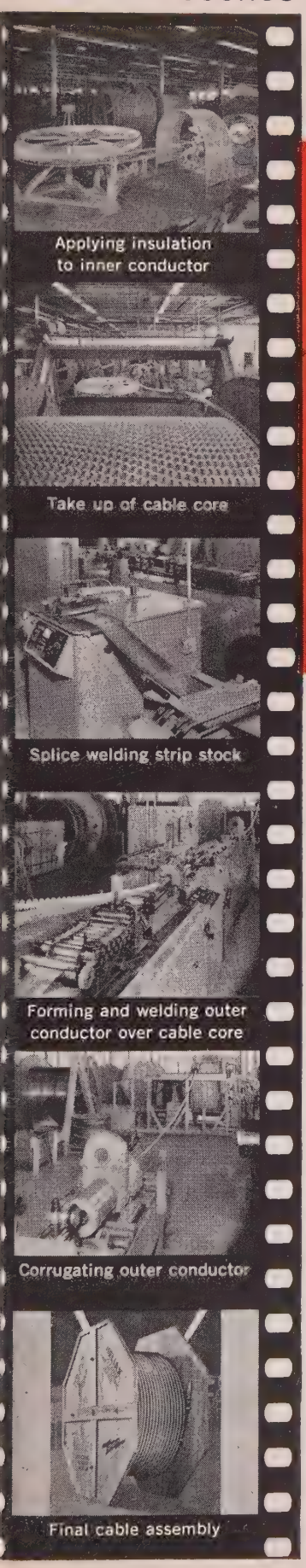
(Continued on page 126A)

ANDREW

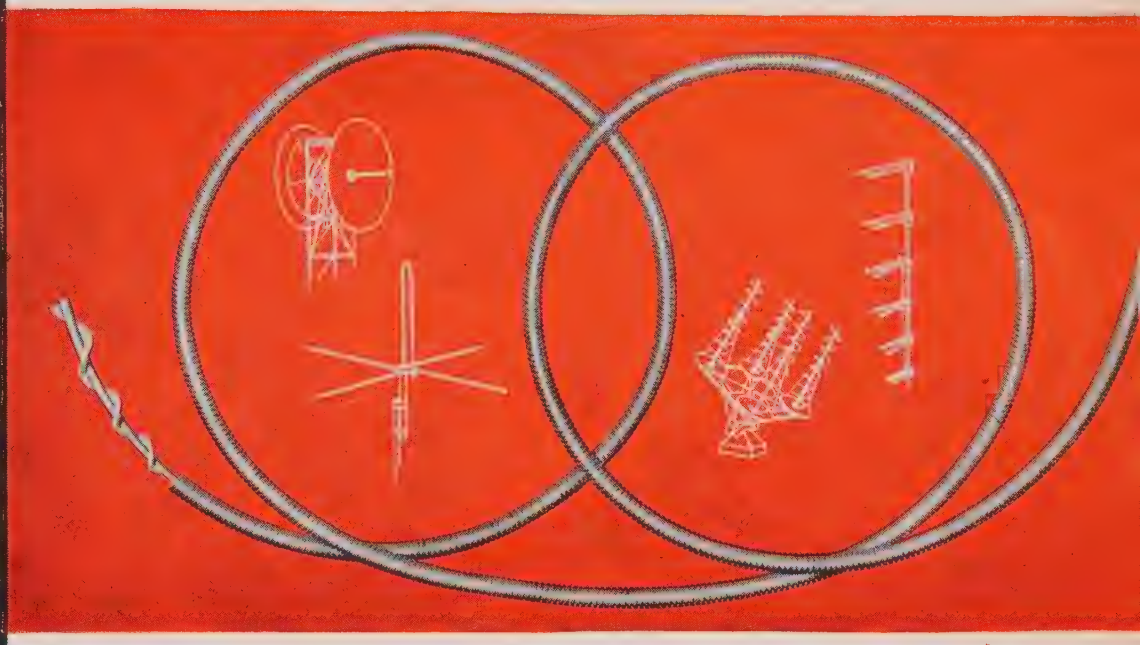
roduction scenes


Applying insulation to inner conductor
Take up of cable core
Splice welding strip stock
Forming and welding outer conductor over cable core
Corrugating outer conductor
Final cable assembly

HELIAX

THE FLEXIBLE AIR DIELECTRIC CABLE

meets every communication requirement



Easiest to install...longest to endure

Produced in the new ANDREW plant facilities, HELIAX is the flexible, low loss, low VSWR coaxial cable for use in all applications from VLF through microwave.

The production cycle in which the copper sheet stock is formed around the cable core, welded and corrugated on a continuous basis is depicted in the film strip. Rolled sheet stock provides considerably closer tolerance tubing than is possible by extrusion or other methods.

HELIAX is the *only flexible* air dielectric cable. This flexibility is imparted by the unique, continuous helical corrugated construction. Bending torque required is about one-half that required for straight wall aluminum or copper tubing of the same size.

HELIAX has the greatest resistance to crushing or kinking. Again this is due to its unique construction. Resistance to physical damage from crushing or kinking forces is about twice that of conventional aluminum or copper cables of comparable size.

HELIAX is the only U. S. produced air dielectric coaxial cable capable of being manufactured in continuous lengths. Critical applications need no longer depend on splicing 1,000 feet or shorter lengths to make up a long cable run. HELIAX affords the only splice free, trouble free air dielectric cable installation.

STANDARD SIZES AND TYPE NUMBERS

SIZE	IMPEDANCE	TYPE NO.
3/8	50	H3-50
7/8	50	H5-50
7/8	75	H5-75
7/8	100	HT5-100
1 5/8	50	H7-50
1 5/8	75	H7-75
1 5/8	100	H7-100
3 1/8	50	H2-50
3 1/8	75	H2-75

ANDREW is your *only* fully integrated source capable of the design and manufacture of cable *and* fittings as well as antennas. ANDREW offers a wealth of engineering experience in the field of RF transmission devices.

You're invited to write for the new HELIAX Catalog H.

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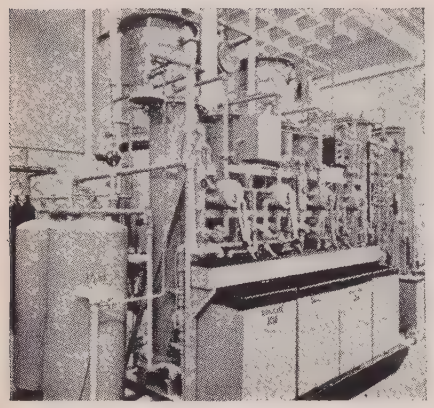
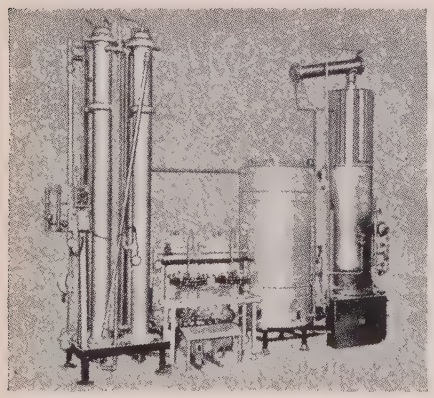
ANTENNAS • ANTENNA SYSTEMS • TRANSMISSION LINES

BARNSTEAD
ENGINEERS
PURE WATER
TO YOUR SPECIFICATIONS



PHILCO'S Lansdale Tube Division uses this Water Purification "Train" equipment in processing components. Costs are reduced as most of the process water is repurified and fed back to the system for re-use. Barnstead equipment includes a 20 GPH High Purity Still, 150 gallon heated ultra-violet equipped tank to prevent growth of bacteria, two BD-10 Holders with special high purity Supercartridges®, an MF-200 Submicron® Filter, and heat exchanger. Barnstead makes water versatile.



IBM installation . . . at Poughkeepsie, New York, produces 200 gallons of extremely pure water per hour, 18 megohms resistance at 25° C. and free of all types of impurities. Equipment consists of Sand and Carbon Filters, Four-Bed Demineralizer, Two High Purity Stills, 1000 gallon storage tank, Mixed-Bed Demineralizer, Submicron Filter, and Heat Exchanger.

HAVE A PURE WATER PROBLEM?
Bring it to Barnstead . . . with over 80 years of experience in all kinds of Water Purification problems. Write for Catalogs; #160 on Barnstead Water Demineralizers, and Catalog "G" on Water Stills.

Barnstead®
STILL AND STERILIZER CO.
56 Lanesville Terrace, Boston 31, Mass.

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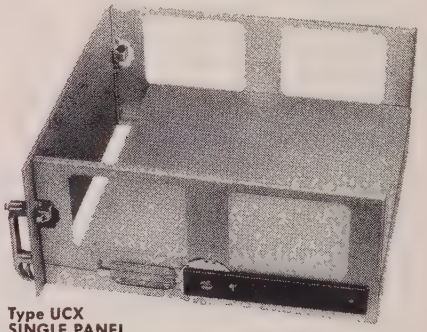
(Continued on page 128A)

RIGID
MOUNTING for YOUR COMPONENTS

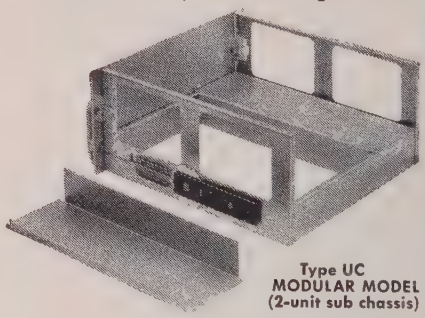
UNIVERSAL CHASSIS

Available in two types: full chassis plate or modular sub chassis.



Type UCX
SINGLE PANEL
CHASSIS PLATE MODEL—with push-button panel lock, trigger 'Detent' tilt control and 'Chassis-Trak' tilt-lock section.

Flat chassis plate is readily punched or drilled, permits fore-and-aft or cross-mounting of assemblies, provides unrestricted front panel mounting area.



Type UC
MODULAR MODEL
(2-unit sub chassis)

This series permits fore-and-aft mounting, superior cooling of assemblies.

• Maximum-height side frames give rigid equipment-protection in any position • Steel or aluminum—1/8" or 3/16" front panels, .062" side frames • All parts drilled for assembly • Panel drilled for handles and chassis drilled for slide mounting ('Chassis-Trak') if desired • Customer may specify all dimensions and materials • Fast delivery

HANDLES:
These features are offered individually or in combination: blank handles, push-button panel locks, trigger for tilt lock, positive screw-down clamp type for extreme shock or vibration.

ONE SOURCE...
for VENTILATED RELAY RACK CABINETS, CONTROL CONSOLES, BLOWERS, CHASSIS, 'CHASSIS-TRAK'*, RELATED COMPONENTS

ORegon 8-7827
WESTERN DEVICES, INC.
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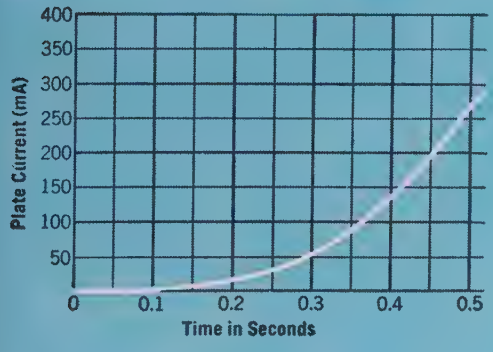
*For Inquiries on 'Chassis-Trak', East of Rockies: Chassis-Trak Inc., 525 S. Webster Ave., Indianapolis

The tube that transistors made necessary!



RF Power Amplifier — Class C Telegraphy

	ICAS 25-50	ICAS 148-174	mc.
Frequency (mobile)			
DC plate voltage	600	600	volts
DC screen grid voltage	200	200	volts
DC plate current	2x40	2x40	mA
DC screen grid current	5.5	4.5	mA
DC control grid current	2x1.2	2x1.3	mA
Plate input	2x24	2x24	watts
Plate dissipation	2x6.5	2x7.2	watts
Power output	35	33.6	watts

Amperex 5895 twin tetrode reaches 85% of full emission in ½ second — minimizes transistor drift.

The Amperex 5895 allows "push-to-talk" operation in compact, transistorized mobile equipment — reaches a practical operating level of 85% of full emission in ½ second (see curve). Minimizes transistor drift and reduces battery drain.

The 5895 RF power amplifier facilitates the design and manufacture of compact, mobile FM VHF/UHF transistorized transmitters.

ask Amperex
for your copy of the latest condensed tube catalog containing data on tubes for mobile operation.

AMPEREX ELECTRONIC CORP. 230 Duffy Avenue, Hicksville, L. I., N. Y.
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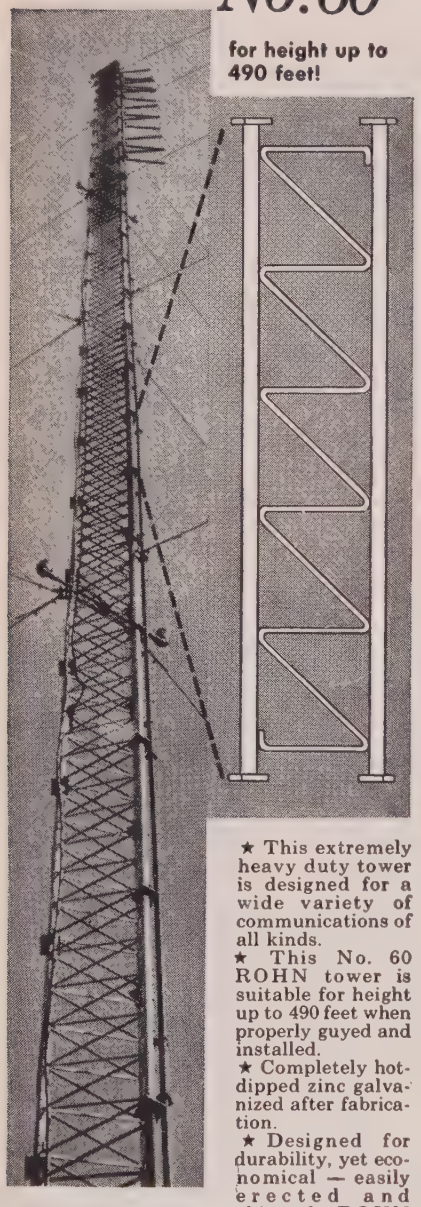
ROHN COMMUNICATION TOWER *No. 60*

for height up to 490 feet!



Shown above is a ROHN No. 60G tower used for FM broadcasting, installed to a height of 300 feet.

★ This extremely heavy duty tower is designed for a wide variety of communications of all kinds.
★ This No. 60 ROHN tower is suitable for height up to 490 feet when properly guyed and installed.
★ Completely hot-dipped zinc galvanized after fabrication.
★ Designed for durability, yet economical — easily erected and shipped. ROHN towers have excellent workmanship, construction and design. Each section is 10 feet in length.

FREE

Details and complete engineering specifications gladly sent on request. Also ROHN representatives are coast-to-coast to assist you.

Write-Phone-Wire Today!

ROHN *Manufacturing Co.*

6718 W. Plank Road, Peoria, Illinois
Phone 7-8416

"Pioneer Manufacturers of Towers of All Kinds"

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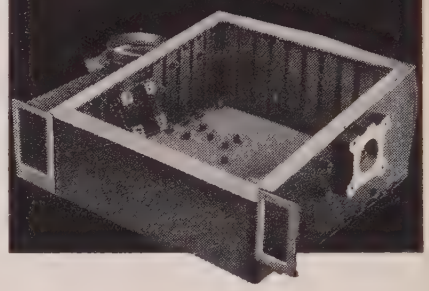
Macurda, C., Manhattan Beach, Calif.

(Continued on page 130A)

Environmental conditioning *for fire control systems*



AiResearch integrated environmental conditioning of aircraft fire control systems achieves high effectiveness while reducing space and weight requirements.

Representative of AiResearch progress is this air-to-air environmental conditioning package which uses the integral heat exchanger and cold plate cooling unit as the base and mounting frame of the pressurized enclosure for the fire control system transmitter. Net weight of this unit is 9.5 lb. with a heat rejection of 425 watts.

This example illustrates the important economies which can be achieved when AiResearch is contacted early in the design stage of electronic systems.

Environmental conditioning equipment has been produced for the following electronic systems:

Detection • Communication • Control • Ground Support • Guidance

Write for literature today.

THE GARRETT CORPORATION

AiResearch Manufacturing Division

Los Angeles 45, California

BUILT-TO-ORDER COMPUTERS

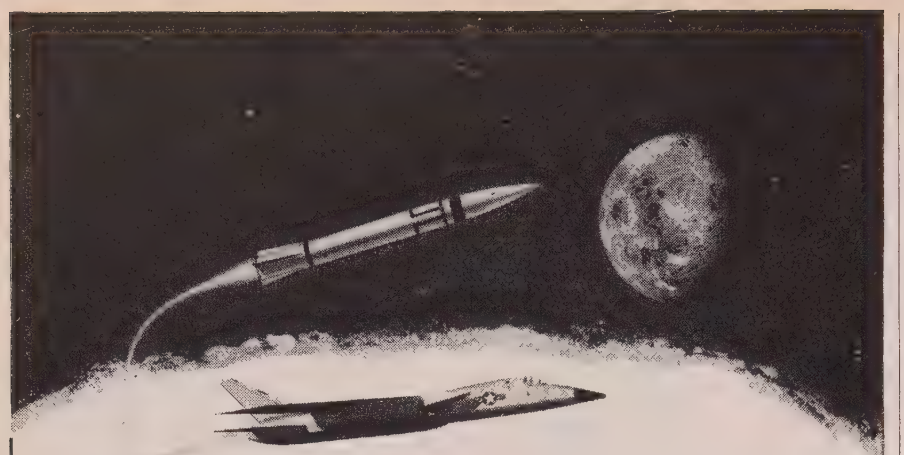
...with off-the-shelf components

Delco Radio can design, develop and deliver digital computers with the speed you need, for airborne guidance and control as well as a wide variety of other special applications. ■ With off-the-shelf Delco transistorized building block modules, we have, for example, built a computer for a military application in less than three months. ■ These miniature modules contain standard components. They satisfy all MIL-E-5272D (ASG) requirements, which assures extremely rugged, reliable computers. Continuing life tests on these computer circuits now exceed four and one-half million transistor hours *without a failure*. And where space is no problem, you can have these same, reliable digital circuits packaged on plug-in circuit cards. ■ Delco Radio has six sections of highly experienced people with the necessary capabilities to produce complete computer systems: Application Analysis, Systems, Logic, Memory, Circuit Design and Advanced Development. May we review your requirements? Just contact our Sales Department. ■ *Physicists and Electronics Engineers: Join Delco Radio's search for new and better products through Solid State Physics.*

PIONEERING ELECTRONIC PRODUCTS THROUGH SOLID STATE PHYSICS

DELCO RADIO

Division of General Motors • Kokomo, Indiana



SYSTEMS ENGINEERS

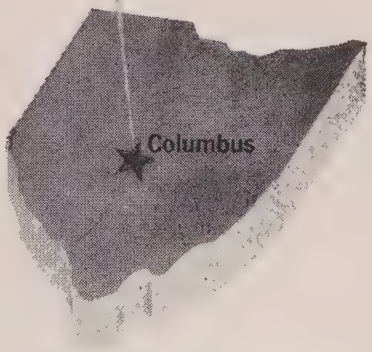
here is your opportunity to join one of the most complete centers of advanced systems capability

The Columbus Division of North American Aviation is a center of *total* systems capability. It is the designer and builder not only of aircraft—such as the A3J Vigilante and the T2J Buckeye—but also of missiles, radio/radar telescope systems, seat ejection systems, and other diverse products. The Columbus Division is also the center of extensive advanced R&D projects. Here, there are unlimited opportunities to contribute to advanced technology—and to forward your own career.

Currently, the Columbus Division has openings for SYSTEMS ENGINEERS. These engineers will assume responsibility for airframe electronic systems. To qualify for these positions, men should have a background in one or more of the following fields: ECM, radar, packaging, reliability, antennas, data processing, reconnaissance or associated systems and components, and design of logic digital computers. They should have a BSEE, or the equivalent, plus three years experience.

If you meet these qualifications, and seek an opportunity to advance your career, please contact us right away. Send a brief resume in confidence to:

Mr. J. A. Hitchcock
Engineering Personnel, Box IR-387
North American Aviation, Inc.
Columbus, Ohio



THE COLUMBUS DIVISION OF NORTH AMERICAN AVIATION, INC.

Membership

(Continued from page 128A)

Maddock, A. W., Willowdale, Ont., Canada
Marshall, T. C., Fullerton, Calif.
McDonald, R. H., Richmond, Ind.
Meeds, L. T., Washington, D. C.
Miley, W. G., Santa Ana, Calif.
Nabel, H. E., Palos Heights, Ill.
Nawracaj, E. P., Chicago, Ill.
Penter, J. L., Sherman Oaks, Calif.
Peterson, F. A., Grand Forks, N. D.
Pritchard, J. R., Titusville, Fla.
Reinschmidt, A. L., Jr., Biloxi, Miss.
Renda, C. J., Youngstown, Ohio
Rockwell, W. D., Brooklyn, N. Y.
Rossi, R. J., Long Beach, Calif.
Sachse, M. H., Los Angeles, Calif.
Schuh, A. G., Concord, Mass.
Shinmachi, T. K., Los Angeles, Calif.
Simmons, W. B., Lexington, Mass.
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Stricklin E. "B," Bainbridge, Md.
Stromberg, R. T., Campbell, Calif.
Stryjewski, A. H., Freeport, L. I., N. Y.
Swedblom, C. E., Ridgecrest, Calif.
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Tominaga, S., Ichikawashi, Chiba-ken, Japan
Werness, H. G., Deerfield, Ill.
Whitehouse, F. E., Severna Park, Md.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

PATENTS OR CONTRACT ADMINISTRATION

Desire to enter field combining the legal and technical. 7 years technical experience (4 years technical writing). Liberal arts degree including 59 credits physical sciences and electronics. Law degree. East coast preference. Resume upon request. Box 2095 W.

ELECTRONICS PATENT LAWYER

Electronics patent lawyer, engineer, etc., 15 years patent experience, desires position handling patent work, and also other responsibilities as time permits, in small dynamic east coast company. Box 2096 W.

(Continued on page 134A)

GLOBAL SYSTEMS ENGINEERING

International Electric is the prime participant — the systems manager — in the development of a global digital command and control system which will provide SAC a rapid, highly modern means of controlling and supporting its forces throughout the world.

Creation of this digital data processing system involves operations research, systems analysis, formulation of equipment requirements, programming and the systems engineering of an integrated complex of electronic sub-systems to meet operational requirements. Electronic systems engineers will find at I.E.C. an exceptional opportunity to express imagination and technical competence.

Engineering positions are open in these areas: systems engineering, logical design, equipment development, preparation and performance of acceptance test, design and application of peripheral equipment. Programmer positions involve creation of advanced techniques in real-time programming, automatic programming simulation and automatic recovery. To inquire about these positions, write Mr. J. J. Crawford, Director of Industrial Relations.

INTERNATIONAL ELECTRIC CORPORATION

An Associate of International Telephone and Telegraph Corporation

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Space Electronics Corporation is actively engaged in both creating and constructing a wide variety of advanced electronic systems vital to continued progress of the nation's missile and space programs. Significant demonstrations of this ability are two highly miniaturized systems conceived and fabricated by SEC for CalTech/NASA Jet Propulsion Laboratory. Consuming less than 1 watt each, they are the Digital Data Automation System and Spacecraft Radio Command System for the advanced lunar and deep space probe known as . . .

Ranger

SPACE ELECTRONICS CORPORATION

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Inquiries from qualified scientists and engineers are invited to the personal attention of Dr. James Fletcher, president.

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Bendix now holds prime responsibility for the ADVENT program for the Signal Corps.

Opportunities are available in:

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Packaging
Environmental Studies
Electromechanical Design
AJ Techniques
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Space Environment
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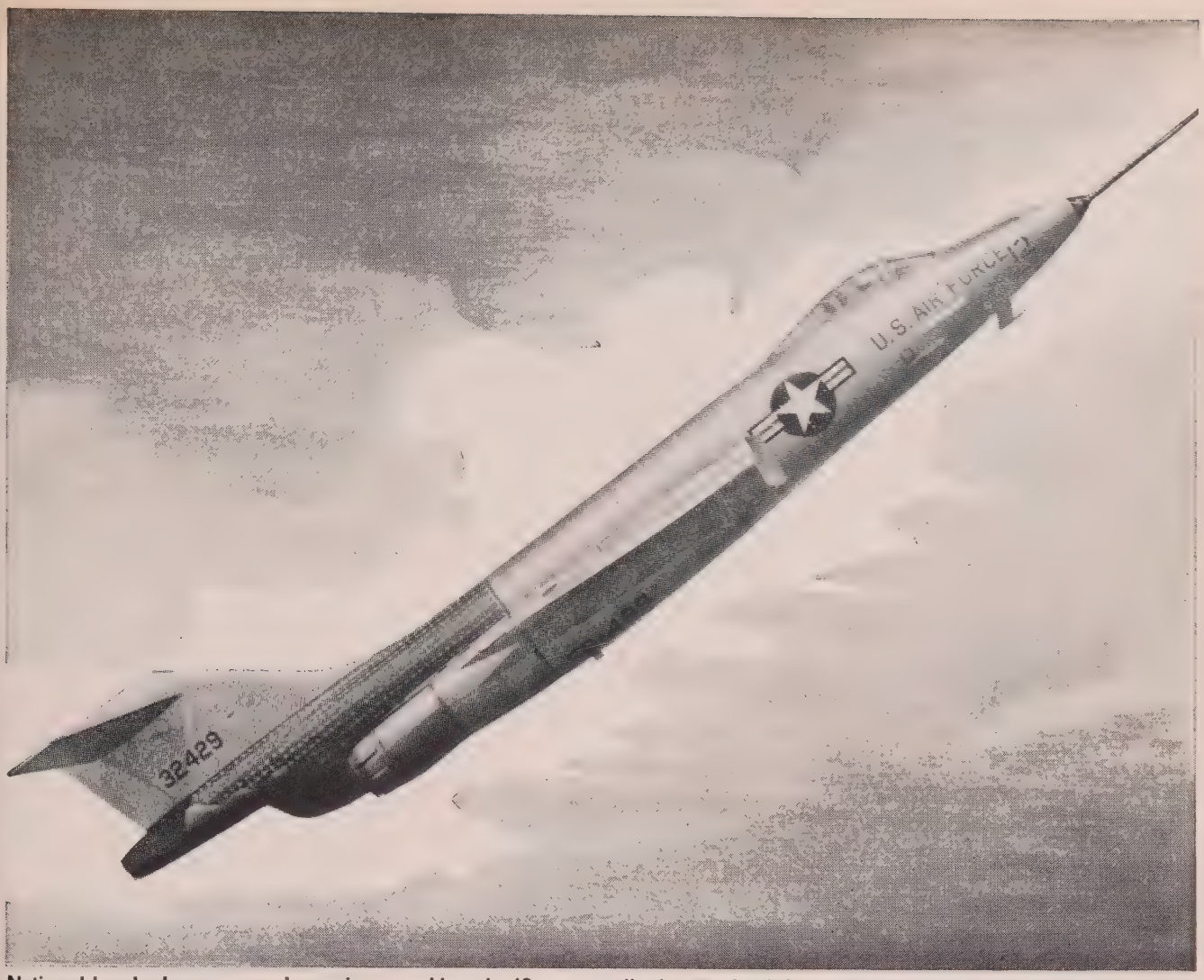
Write today to Director of Personnel, Dept. L11.

THE Bendix CORPORATION

BENDIX SYSTEMS DIVISION

ANN ARBOR, MICHIGAN

(F101, AIR FORCE PHOTO)



National has had many years' experience making significant contributions to the defense effort, including airborne components.

COMPARE THIS OPPORTUNITY
WITH WHAT YOU ARE NOW DOING!

HERE'S WHAT *National's* NEW MILITARY RESEARCH AND DEVELOPMENT PROGRAM OFFERS.

This operation will interest any engineer or scientist possessing enough self-confidence—ability and experience—to develop projects initially and carry them through to completion.

WHO WE'RE LOOKING FOR

National is looking for military-oriented scientists and engineers who hold a B.S. degree or advanced degrees. You should be working in electronic, electro-mechanical, mechanical, physics, optics, mathematics, or other related areas. Preference will be given to those who have had several years' experience dealing with prime contractors and government agencies.

As a member of National's New Military Development Team—you will be working initially with our Military Proposal Group. As proposals become specific projects, your responsibility will continue through the contractual stage for technical liaison, fulfillment of contractual obligations including hardware development, meanwhile retaining sufficient flexibility to continue your proposal efforts.

WHY YOU SHOULD INVESTIGATE

National's new Military Research and Development Program offers you unusual latitude in responsibility. It offers you the chance to participate in military projects from start to finish. Furthermore, you now have the opportunity to join an operation still in its formative stage—yet backed by one of the world's most successful . . . most reputable corporations.

COMPLETE INFORMATION is yours by sending your résumé to Mr. T. F. Wade, Technical Placement Section F2-5, The National Cash Register Company, Dayton 9, Ohio. All correspondence will be kept strictly confidential.

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ENGINEERS

WHY SANDERS PROVIDES REAL OPPORTUNITY

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- From 11 men to 1400 in only 9 years—and still growing
- Gross sales increasing steadily—$17 million in fiscal year 1960—with 50% increase projected for 1961

This means new opportunities year by year for forward-looking, ambitious engineers

STABILITY

- Backlog of $50 million at time of writing
- Most of the company's developmental work and specific products are in areas of the electronics industry where overall growth outlook is good—military systems, automatic assembly and microwaves' instrumentation and control

Long-term assignments assured on a diversity of contracts

REPUTATION

- PANAR® radar, FLEXPRINT® flexible printed circuits and TRI-PLATE® microwave products paved the way for new prime contracts—$30 million worth, primarily for production of electronic systems that have emerged from the company's research and development programs.

Increase your own knowledge and professional stature by working with men who rely on new technical concepts rather than traditional approaches

INQUIRE NOW ABOUT POSITIONS IN:

CIRCUIT DESIGN—SENIOR LEVEL
Particular emphasis on transistor application to analog and digital techniques; data handling equipment; audio, video, RF circuitry and switching.

SYSTEMS
ECM, Radar and ASW. For theoretical studies and hardware implementation. Through Project Engineer level.

PACKAGING
For both airborne and ground based equipment. Experience in vibration, heat transfer, printed circuit techniques.

TRANSMITTER MODULATOR DESIGN
To contribute to the design of the Eagle Missile transmitter modulator. Experience with modulator or transmitter circuits and system design of oil immersed and dry designs on light weight units of ultra reliability. General experience with vacuum tubes and pulse circuits, high voltage design and magnetic circuits.

TEST EQUIPMENT
To design and develop test equipment for both manufacturing and research and development activities. Good knowledge of pulse and low frequency circuits, switch measurements, sweep circuits and the principles of pulse doppler radar.

RECEIVER DESIGN
VHF electronically scanned airborne receivers, filters, problems in spurious response reduction and multiplexing.

Please address your inquiries to
Mr. Roland Hood, Employment Manager

SANDERS ASSOCIATES, INC.

NASHUA, NEW HAMPSHIRE
(less than 1 hour from downtown Boston)

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Positions Wanted

(Continued from page 130A)

SEMICONDUCTORS

Age 34. 7 years background engineering supervisor and factory superintendent, transistor diode and rectifier production. B.S. in chemistry, graduate work and management training. Earlier experience quartz oscillators. Desires more responsibility. Box 2097 W.

BASIC RESEARCH

BSEE. (Electronics option). University of Michigan 1948. Age 36. Member IRE. Wide industrial and some college teaching experience in electronics. Have a great desire to do basic research in such fields as atom structure and atom particles, propagation of light energy, magnetism, and gravitation. Seek college research assistantship or fellowship for the 1960-1961 academic year. Box 2098 W.

TEACHING

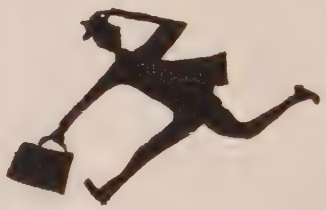
Ph.D. Engineer-Physicist; age 38; Member IRE; American Phys. Society. Experience: 16 years teaching, research, industry, including supervisory. Publications. Specialty microwaves and devices (PGMTT, PGED). Seeks full professorship or chairmanship in reputable, growing E.E. Dept. Box 3001 W.

ELECTRONIC ENGINEER

MEE. Seeks affiliation with, or will organize planning staff for technical-economic evaluation

(Continued on page 138A)

Are You "Going Someplace?"



Check These FOUR Top-Notch Opportunities for Men on the Way Up!

Expansion has created opening for a *Mkting. Mgr.* who will be expected to move up to *VP* in a short time. Also needed at once are a *Sales Manager* with telecommunications or R&D background and *2 Sr. Sales Engrs.* Compensation for all positions is excellent.

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Professional placement report
for Electronics Engineers

How the continuous need to improve the nation's space surveillance capabilities opens avenues for new engineering careers

The continuous need we are talking about at General Electric refers to the fact that future-generation missiles, satellites and deep space probes will require refined or entirely new detection techniques, including many that have not yet been conceived.

For example, it is anticipated that for every technical discipline now utilized in the detection field, at least one more must be found to apply within the next 10-year period.

With this in mind, General Electric is increasing its electronics engineering staff now working on advanced missile, satellite and deep-space-probe detection systems. Keeping pace with this expansion, the Company added a new building last year, and another will be ready for occupancy in a few months.

First clues to this trend were obvious in General Electric's well known "Golfball Study," published five years ago. This study compared the problem of missile detection to that of locating a golfball 200 miles away, using the most advanced techniques available in 1954. The problem no longer has such proportions, thanks to the creative imagination of dedicated General Electric engineering, scientific, and technical personnel responsible for designing and developing the unique surveillance sub-system of the Ballistic Missiles Early Warning System (BMEWS) which is receiving headline attention today.

Find out more about these creative and self-expressive opportunities now open to qualified personnel in one of the most vital technologies of the space age.

GENERAL ELECTRIC

Immediate openings for qualified electronics engineers

RADAR EQUIPMENT SYSTEMS SPECIALISTS capable of conceiving and directing the design of long-range radar systems. Desirable experience includes 3 to 10 years in at least one of the following: radar systems design, antenna systems, RF components, transmitters, radar receiver systems or radar data processing systems. Salary structure is fully equal to the professional requirements of the job.

ADVANCED SYSTEMS ENGINEERS capable of defining future defense and space detection problems including deep space-probe tracking. Also the ability to conceive and establish the feasibility of optimum systems solutions to these problems—making use of the most advanced techniques and understandings. Also required is an ability to recognize the need for, and coordinate the development of, new techniques and the exploration of new phenomena. Experience requirements include a Bachelor degree plus a combination of advanced training and several years' experience in both the theoretical and practical aspects of detection systems engineering. A desire to work in the conceptual phase of systems design with the analytical ability required to evaluate and demonstrate the effectiveness of the proposed systems is essential.

FIELD OPERATIONS ENGINEERS for systems management teams to be deployed at complex radar systems installations of the BMEWS type. Systems-oriented Electronics Engineers are needed who have the ability to assume responsibility for installation, checkout, and integration of major radar systems. Background in high-powered Klystron transmitters, low-noise receivers and digital data processing equipment is desirable. A Bachelor degree is required.

ADDRESS YOUR INQUIRY IN CONFIDENCE TO:
MR. JOSEPH WOOL, PROFESSIONAL PLACEMENT, DEPT. T-9
MISSILE DETECTION SYSTEMS SECTION
HEAVY MILITARY ELECTRONICS DEPARTMENT
GENERAL ELECTRIC COMPANY
COURT STREET, SYRACUSE, NEW YORK

Qualified applicants will be invited to visit us in Syracuse at Company expense. Relocation assistance will be provided. 177-25

GENERAL ELECTRIC

Key positions on new semiconductor projects with IBM Research

SEMICONDUCTOR DEVICE AND CIRCUIT STUDIES

Positions are now open on new semiconductor device and circuit studies being undertaken at the IBM Research Center at Yorktown Heights, New York. These studies will present unique opportunities to do advanced work in your field.

Advanced degree in E.E. or equivalent desired to engage in advanced semiconductor device measurement and analysis. Special emphasis will be placed on the frequency range between 500 and 5000 mc/sec. Prefer some acquaintance with semiconductor physics. The work will entail cooperation with outstanding semiconductor scientists at the IBM Yorktown Laboratory in evaluating the physical mechanism involved in newer devices such as the tunnel diode.

Advanced degree or equivalent in E.E. or Physics to engage in semiconductor circuit analysis. Special emphasis on computer application of high-speed semiconductor devices including transistors and tunnel diodes. Opportunity to examine quantitatively the merits of particular devices and to formmulate criteria for comparative performance. The work will be done in close cooperation with device fabrication scientists. Ultimate aims include consideration of integrated or preconnected computing circuits wherein the transmission lines are treated as integral parts of the device.

IBM's ultra-modern new research center is located in one of the most desirable parts of Westchester County, just forty miles north of New York City. All IBM benefits apply. Your relocation expenses paid.

For further information, please write, outlining your background and experience, to:

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A PLACE FOR YOU awaits at rapidly expanding Philco Western Development Laboratories on the San Francisco Peninsula, where new concepts of space communication offer exceptional opportunities to the exceptional engineer . . . with proportionate rewards. Your experience and your interests determine your assignment at PHILCO in the development of the complete system — from the space communication network of space vehicles to the complex data processing equipment of the ground station complex.

For you . . . the opportunity of graduate study on a Tuition Refund basis at any of the excellent surrounding educational institutions, liberal employee benefits, and the facilities of Philco's new, modern R & D laboratories.

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Immediate openings now exist for graduate engineers with experience in equipment design, systems engineering, analysis and integration. As a first step toward expanding your skills at Philco, send your resumé in strict confidence to Mr. W. E. Daly, Engineering Placement.

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Positions Wanted

By Armed Forces Veterans

(Continued from page 134A)

of new ideas, products and acquisitions in the electronic and electromechanical fields. 15 years diversified experience R&D, studies, project engineering, proposals and presentations, planning and requirements. Top references. Box 3002 W.

ELECTRONIC ENGINEER

Electronic Engineer 1943 B.S. Extensive laboratory and supervisory experience in R&D and applications of electron tubes and solid-state devices in military and industrial equipment. Desires responsible technical position in Australia, preferably with American affiliated company. Box 3003 W.

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Industrial, technical, or vocational teaching of radio, TV, and Industrial Electronics. B.S. in Television Engineering. Anywhere in the U.S. Age 40. Box 3004 W. 10 years experience.

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Data Communication Systems. Regional sales and field engineering manager, military and commercial accounts. Standard and special systems using wire line, radio and microwave transmission. Also experience in computer and special purpose EDP systems sales, advertising and sales promotion. BSEE. Age 37. 13 years experience. Box 3007 W.

(Continued on page 140A)

MICROWAVE ENGINEER

THE RADIATION LABORATORY OF THE JOHNS HOPKINS UNIVERSITY HAS A POSITION FOR A QUALIFIED ENGINEER IN THE FIELD OF:

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THE JOHNS HOPKINS UNIVERSITY

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Portrait of a Mach number

Air blasting across an aerodynamic shape at Mach 2 (above, left) records its image on film. In private industry's most extensive complex of wind tunnel installations, Boeing engineers and scientists are defining the shape of the future in supersonic and hypersonic flight. A new hypersonic tunnel, the nation's largest privately owned facility of its kind, tests up to Mach 27.

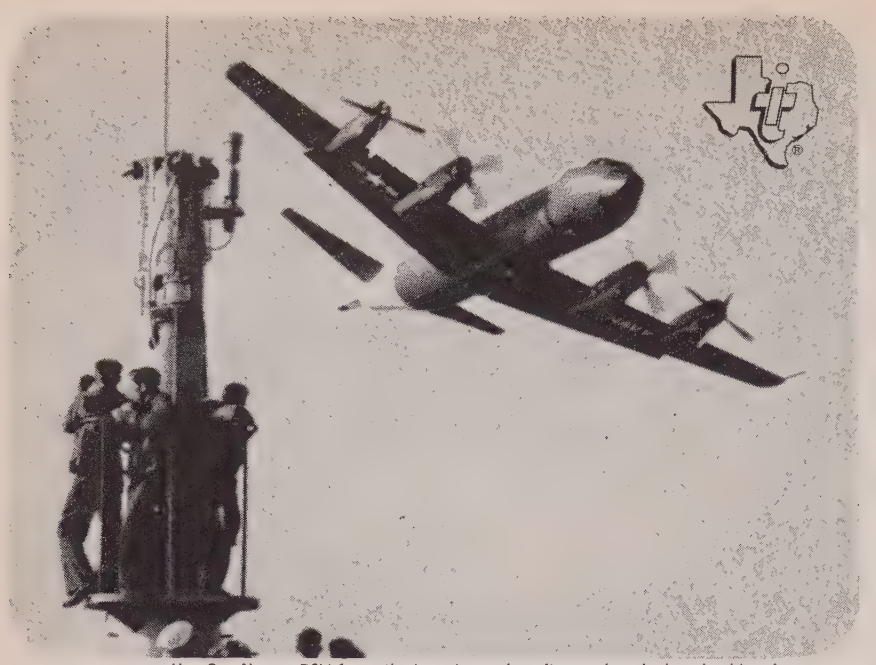
Boeing's emphasis on research and development of future advances covers a wide variety of fields, including missiles, satellites, space vehicles, anti-submarine warfare systems, hydrofoils, commercial and military aircraft, gas turbine engines, electronics, communication, propulsion systems, vertical and short take-off and landing aircraft.

Professional-Level Openings

Expansion of advanced projects and systems management programs of the future has created openings at Boeing for professional specialists in scientific and engineering disciplines, and other, non-technical, areas of company activity. You'll find at Boeing a professional environment conducive to deeply rewarding achievement. Drop a note, mentioning degrees and major, to Mr. John C. Sanders, Boeing Airplane Company, P. O. Box 3822-PIJ, Seattle 24, Washington.

BOEING

Divisions: Aero Space • Transport • Wichita • Industrial Products • Vertol • Also, Boeing Scientific Research Laboratories



U. S. Navy P3V-1 antisubmarine aircraft produced by Lockheed — equipped with TI-built AN/APS-80 surface search radar, AN/APA-125A indicator, and additional gear (nomenclature classified).

AIRBORNE SUB-HUNTER USES TI ASW SYSTEMS

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SYSTEMS:

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Massachusetts Institute of Technology
BOX 16 • LEXINGTON 73, MASSACHUSETTS

Positions Wanted

By Armed Forces Veterans

(Continued from page 138A)

ELECTRONIC SALES EXECUTIVE

South or Central America desired, as representative for electronics component or system manufacturing. BSEE, 10 years applied experience from Semiconductors to Systems. Demonstrated managerial ability. Fluent in Spanish and other languages. Married and family. Previous overseas engineering experience. Member of IRE, age 37. Box 3008 W.

ELECTRONICS ENGINEER

BEE, MSEE, 1st Lt. Army Ordnance. INDUSTRIAL EXPERIENCE: 17 months as research engineer, military communication systems, programming digital computers. MILITARY EXPERIENCE: 2 ¾ years to include two new missile systems as technical and project officer. Available April 1961. Box 3009 W.

ENGINEERING EXECUTIVE

15 years experience in directing engineering departments in field of entertainment electronics, with additional background in manufacturing, purchasing, finance, and general management. Expert in handling growing pains of small company in $2,000,000-$5,000,000 yearly gross sales category. Age 40. Will consider participation. Box 3010 W.

ENGINEERING TECHNICIAN

Graduate of American Television Institute of Technology, with B.S. in Television Engineering. Married, age 37. 10 years experience in Radio and TV Broadcast Engineering. Wants position with up and coming firm that is going places with opportunity for further study and advancement desirable. Box 3011 W.

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

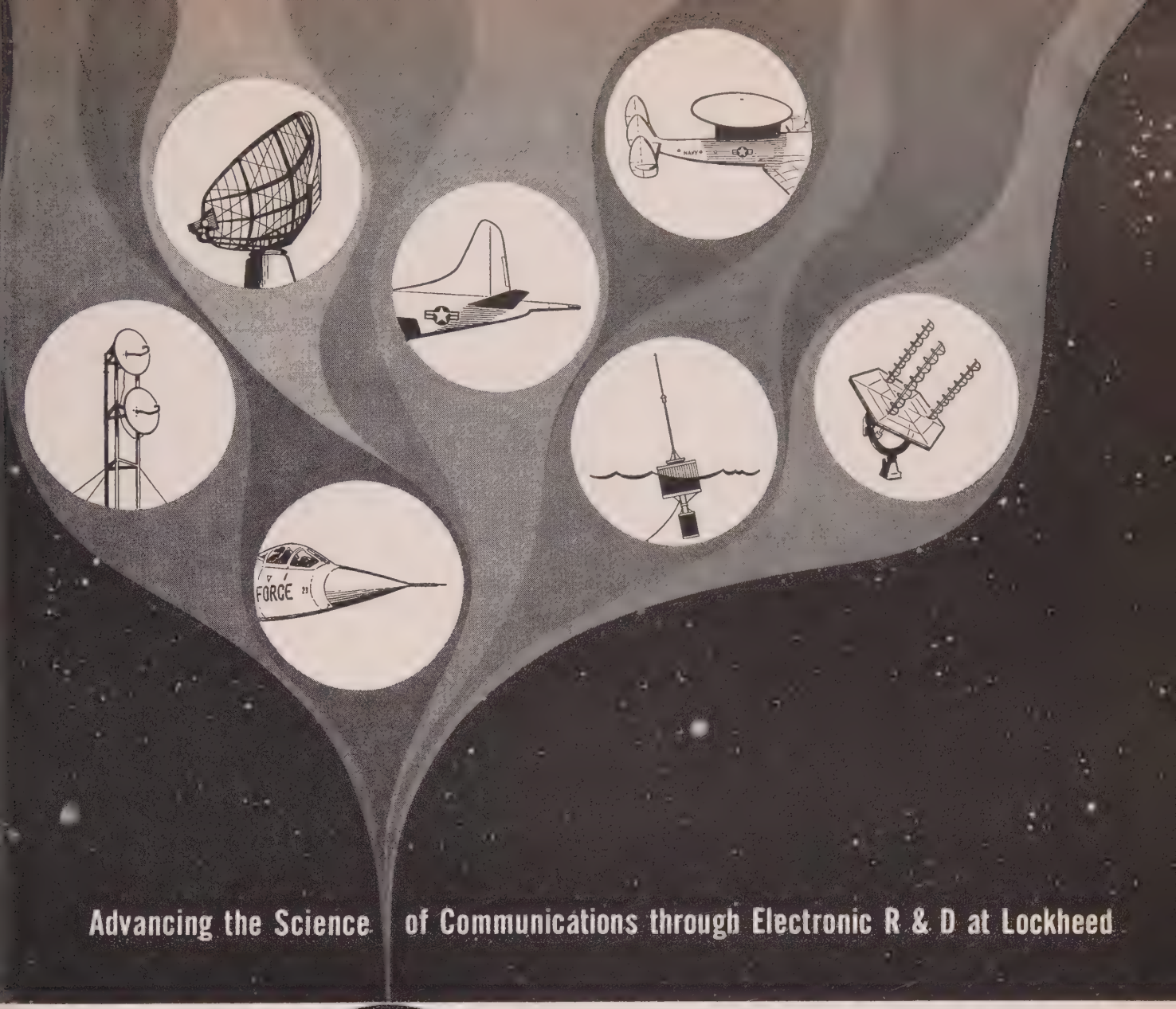
The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

RADIO AND TELEPHONE ENGINEERS

Hawaiian Telephone, serving all of the 50th State, has positions available for experienced radio and telephone engineers. Our rapid growth requires expansion of VHF and microwave communications systems. Experience is desired in the design of radio systems to telephone company standards and in the preparation of associated FCC applications. Openings are at several levels and salary will be commensurate with training and experience. Please airmail resume, with photo, to Employment Supervisor, Hawaiian Telephone, P.O. Box 2200, Honolulu, Hawaii.

(Continued on page 142A)



Advancing the Science of Communications through Electronic R & D at Lockheed

The science of communications is vast as the universe itself. Fresh areas in this science are being explored at Lockheed. These cover the spectrum of communications problems—on and under water to tracking missiles and satellites—from components to complete systems.

An essential phase in the electronics communications R & D program is the development of antennas and supporting equipment to receive telemetered, tracking and relay data. Each is designed to meet a specific need. This program is vital to our sophisticated spacecraft projects—now and in the future.

Areas under investigation in electronics and other fields stir the creative mind: Design and development of data processing equipment; V/STOL design and development; electromagnetic research in corona and high-altitude breakdown studies, surface wave generation, antenna vehicle interaction, millimeter wave radiometry; electrical instrumentation; infrared and solid state physics; biophysics research (on radiation hazards coincident with space flight); solid state electronics; aerothermodynamics; dynamics; autocontrols; and servosystems.

Scientists and Engineers: If you are experienced in any of the areas mentioned above; if you are interested in joining a company that looks far into the future, you are invited to investigate the opportunities offered. Write today to: Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1811, 2402 No. Hollywood Way, Burbank, California.

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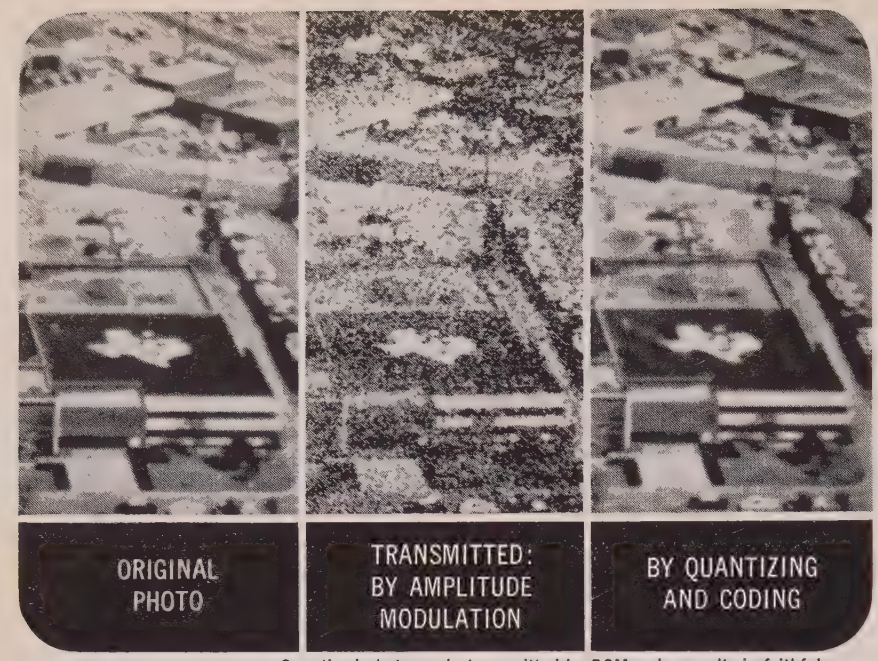

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TI INNOVATION IN SURVEILLANCE SYSTEMS

APPARATUS DIVISION

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Professional Opportunities Are Available For

Electrical Engineers

with interest and experience in the following fields:

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- Operation & Maintenance of Nuclear Devices

For information please write to:

Personnel Manager

Brookhaven National Laboratory

UPTON, LONG ISLAND, N. Y.

BROOKHAVEN NATIONAL LABORATORY
A Center for Nuclear Research and Development
UPTON-NEW YORK

Positions Open

(Continued from page 140A)

MEDICAL ELECTRONICS

A recent graduate with background of practical experience wanted to organize and supervise a medical electronics service in a rapidly expanding medical school. Contact Dr. James Randall, University of Missouri Medical Center, Columbia, Missouri.

ENGINEERS—TEACHING

Expanding urban engineering school in upstate New York requires Ph.D.'s or equivalent for undergraduate and graduate teaching and research in electrical engineering in the areas of computers, advanced electronics, communications, solid-state, and control engineering. Liberal policies, excellent salaries and benefits. Box 2028.

CONSULTING ENGINEER

Well-known Washington Consulting Engineer is seeking a congenial partner or associate who feels that he too needs to expand by associating with co-consultants to widen their mutual operations. Naturally, such a man should be an experienced Consulting Engineer able to contribute his own background and training to the partnership. On the other hand a capable young man with the requisite technical training and interest and abilities who has not yet achieved professional standing but with proper association is ready for registration could buy into the partnership and grow with it, if he is really dedicated to broadcasting. Box 2029.

(Continued on page 144A)

IF YOU'RE AIMING FOR THE TOP . . .

Let Abbott's be your Personal Agent in presenting your qualifications to the nation's foremost employers of engineers and scientists. You'll find this specialized service extremely effective in placing you where your ability, background and effort will yield the greatest rewards.

If you are placed through us, our clients will pay all expenses and our service charges.

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OPERATIONS MANAGER—Digital Computers
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SYSTEMS ENGINEERS—Radar, microwave, telemetry, information and communication theory
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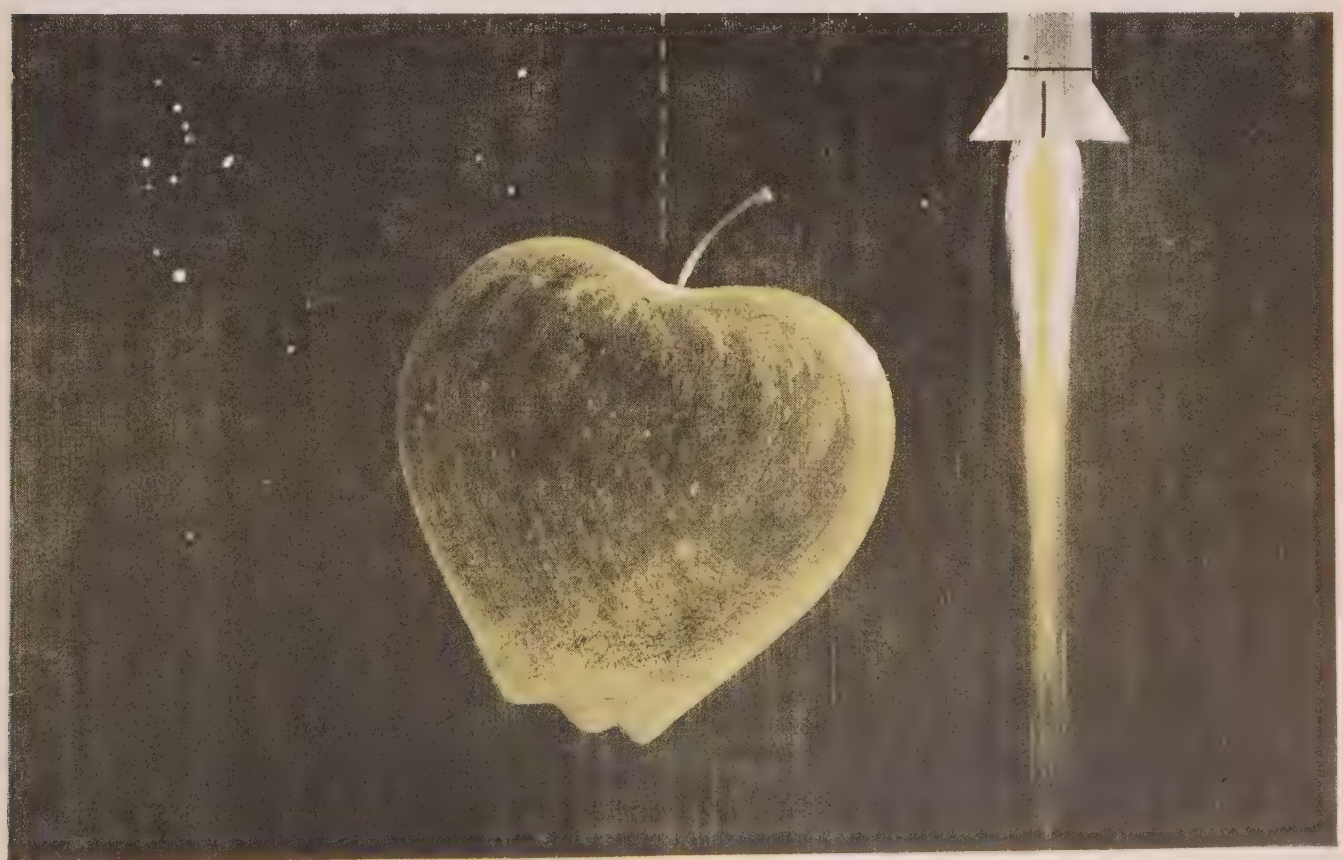
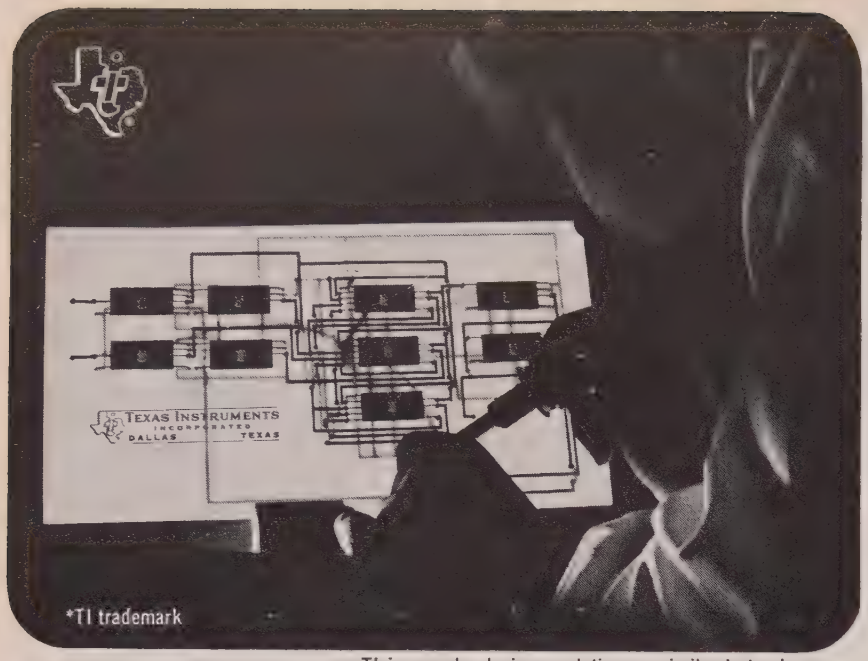
BUT, SIR ISAAC, WE'RE NOT GUIDING APPLES!

Until a ballistic missile is in free fall, our inertial guidance systems must be able to account for both rocket thrust and gravity. Making them this smart is a tough job, but we hit the mark so well on Thor that all of this country's long range missiles will soon be guided inertially. If you would like to help us keep pioneering new guidance systems, and have a BS, MS or PhD in Physics or Math, or an ME or EE, please contact Mr. B. B. Allen, Director of Scientific and Professional Employment, 7929 S. Howell Ave., Milwaukee 1, Wisconsin.

AC SPARK PLUG ⚛ THE ELECTRONICS DIVISION OF GENERAL MOTORS





TI is now developing revolutionary missile electronics using SOLID CIRCUIT semiconductor networks.

SOLID CIRCUIT* NETWORKS IN TI MISSILE SYSTEMS

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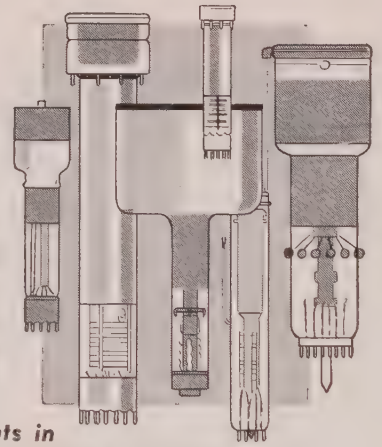
Outstanding Opportunities for

ENGINEERS PHYSICISTS



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- DESIGN
- DEVELOPMENT
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- SPECIAL PURPOSE TUBES
- IMAGE TUBES
- POWER TUBES
- MICROWAVE TUBES
- SPECIAL ELECTRON DEVICES

Write or send resume to:
Mr. Wm. Kacala, Technical Recruiting
Box 284, Dept. M-E42, Elmira, N.Y.
or phone collect Elmira REgent 9-3611

Westinghouse
ELECTRONIC TUBE DIVISION ELMIRA, NEW YORK

Positions Open

(Continued from page 142A)

CONSULTING ENGINEER

Washington Broadcasting Consulting Engineer has vacancy for assistant on full or part-time to do drafting and engineering work, and similar work in connection with broadcast station applications. Some previous experience in this field would be desirable, but a beginner with an adequate engineering background would be acceptable. Box 2030.

ELECTRONICS SECTION HEAD

An attractive opening is available as technical supervisor of a 10 man section of experimentally oriented electronics engineers and technicians. Position holds outstanding opportunities for growth from an already well established research program, both by way of personnel and highly successful technical projects. Horizons not limited by marketing and sales. Experience in experimental radar, radar data processing, or computers, is necessary to command the respect of and provide the leadership for personnel already in the section. The Lab. provides a desirable combination of the best features of industry as to financial and equipment resources coupled with a university associated atmosphere for constructive R and D activity at the highest level. Salary range starts at $13,000 depending on qualifications and experience. Send complete resume to J. L. Rentschler, Cornell Aeronautical Lab. Inc., 4455 Genesee St., Buffalo 21, N. Y.

(Continued on page 148A)

ELECTRONICS ENGINEERS AND PHYSICISTS

B.S. or M.S. in E.E. with electronics major and at least 4 years experience in 3 of the following fields: Electronic Circuit Design, Wave Shaping and Timing Equipment, Electronic Digital and Anulog Computation or Servo Theory.

Ph.D. in physics or electronics. No experience necessary but prefer candidates with interest in specializing in electron optics and magneto-hydro-dynamics.

These are permanent positions offering good salaries and opportunities for personal and professional growth. Liberal benefits include well-established pension, group insurance and stock investment plans. Our modern laboratories are located in a pleasant suburban location in north central New Jersey 45 minutes from New York City. Mail complete resume with salary requirements to personnel manager.

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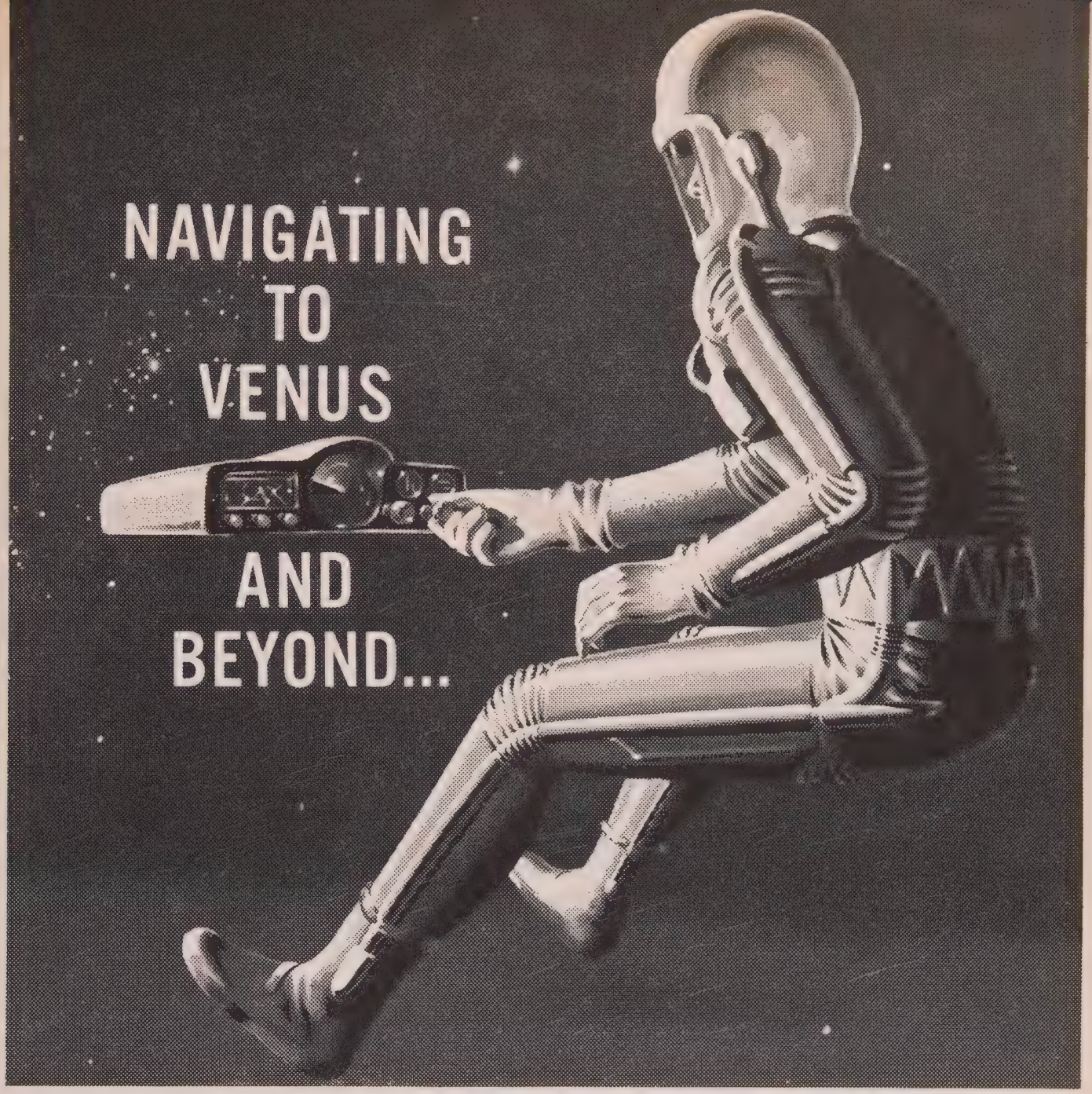


A CHALLENGE TO EXCITE ANY ELECTRONICS ENGINEER WORTH HIS SALT...THE KIND RYAN NEEDS RIGHT NOW

We do not know where Venus is, within 50,000 miles or more. What is needed is a guidance system capable of injecting space vehicles into interplanetary orbits with the accuracy required for advanced space missions.

We are confident that the C-W doppler systems, of which Ryan is the world leader, will meet this challenge of space navigation. This is one reason Ryan is the largest electronics firm in San Diego—*and the fastest growing!* If you are an electronics engineer ambitious to help advance the art, as well as your own career, through vital "frontier work"—we want you *right now* at Ryan Electronics.

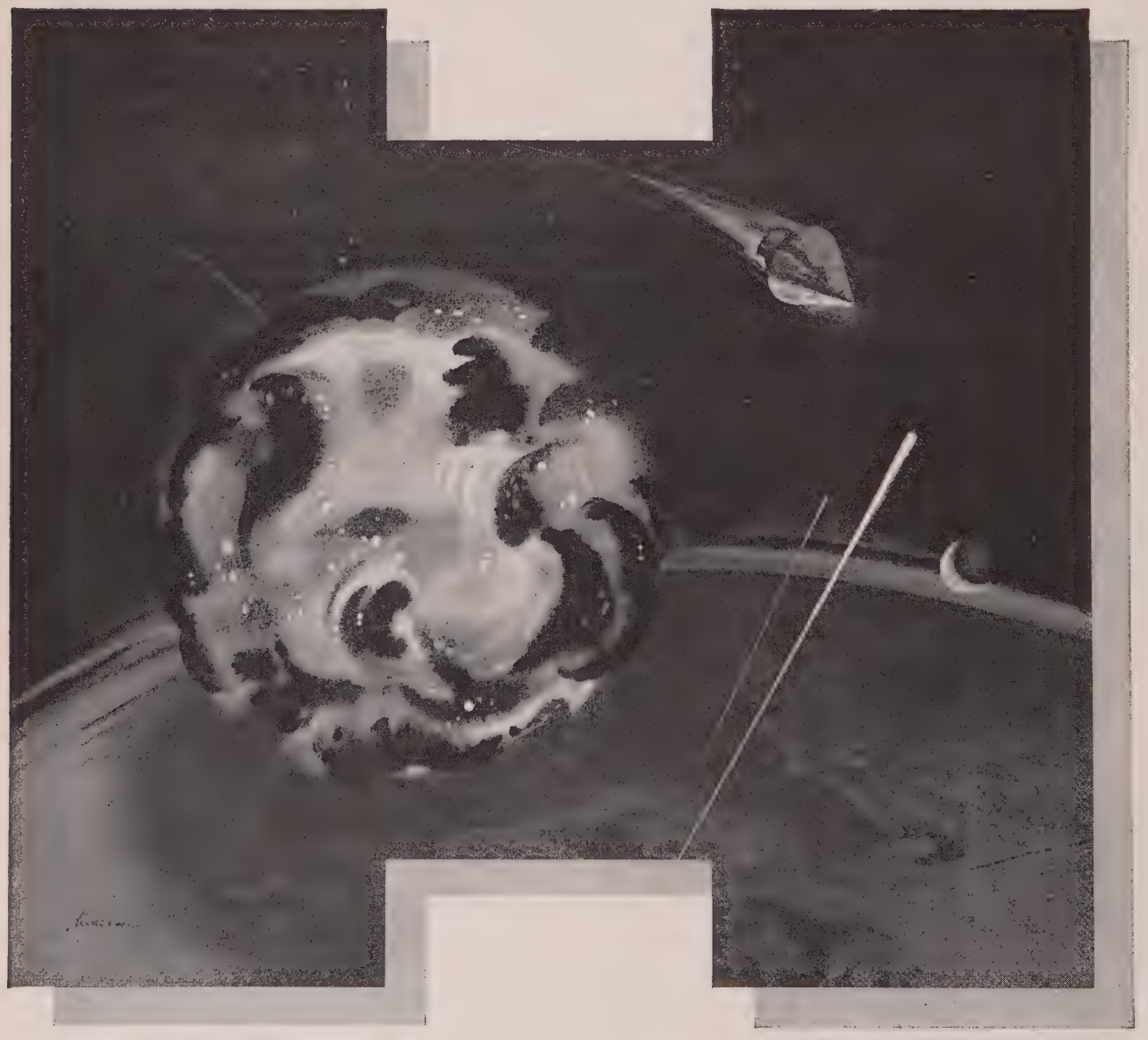
Ryan Electronics employs over 2000 people and has over one-third of the company's $149-million backlog of business. Under the leadership of some of America's most prominent scientists and engineers, Ryan is probing beyond the known . . . seeking solutions to vital problems of space navigation.

Expanding facilities of Ryan Electronics at San Diego and Torrance in Southern California are among the most modern in the West. You enjoy living that's envied everywhere, plus facilities for advanced study. Send your resume or write for brochure today: Ryan Electronics, Dept. 1, 5650 Kearny Mesa Road, San Diego 11, Calif.

DIVISION OF RYAN AERONAUTICAL COMPANY
SAN DIEGO & TORRANCE · CALIFORNIA

RYAN ELECTRONICS

Ballistic missiles: the ultimate



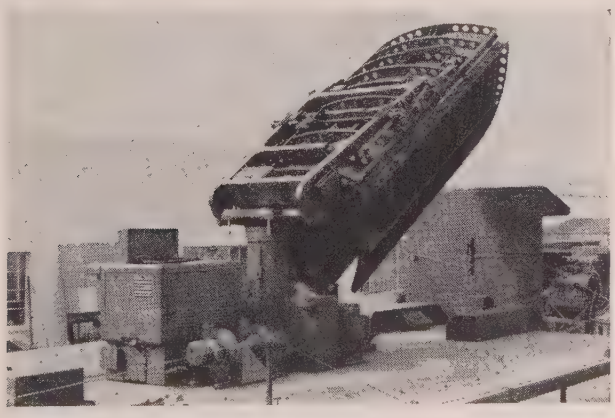
weapon?

The ICBM is often called the "ultimate weapon." Yet, throughout history, there have been many so-called ultimate weapons. Men of science have always found a defense.

Today, Hughes engineers are studying this problem. The programs being initiated are uncovering many challenges for imaginative engineers and scientists.

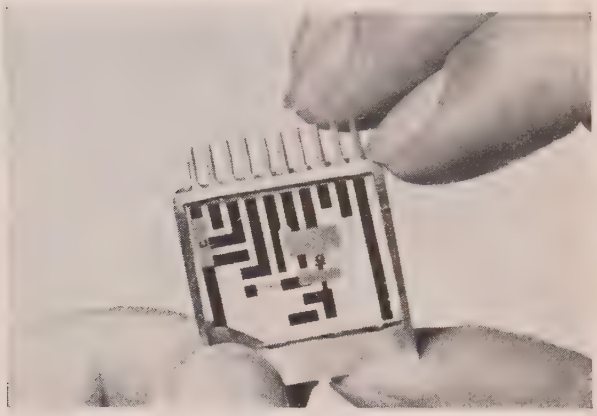
The high order of complexity in ballistic missile interception means that all of the traditional areas involved—early detection, discrimination, acquisition, tracking, data processing, intercept and kill—must be expanded to new highs in the state of the art.

Of tremendous help are the lessons learned during the development of the first successful airborne electronics intercept system—made up of the Hughes Electronic Armament System and Hughes Falcon air-to-air-guided missiles.



This Hughes three dimensional radar antenna, *shown in the process of retraction, is completely mobile. Developed by Hughes engineers, frequency-scanning radar provides umbrella-like protection for key defense areas.*

This complete flip-flop circuit *demonstrates Hughes capabilities in the field of solid state data processors. Hughes engineers are broadening the state of the AICBM art by drawing upon a great number of such advanced electronic techniques.*



Hughes is also drawing on its skills in such fields as advanced infrared and radar detection systems, satellite communications systems, electronic scanning radar systems, space vehicle technology, microwave propagation, advanced data handling and display systems, and other fields of advanced electronics.

Hughes Anti-ICBM work will be coordinated to combine the efforts of virtually every Hughes activity. In this way, every conceivable idea can be studied—and the entire spectrum of electronics broadened.

The great variety of advanced electronics projects and the basic scientific-orientation at Hughes offers the engineer the ideal environment for personal and professional growth.

Newly instituted programs at Hughes have created immediate openings for engineers experienced in the following areas:

Computer Components	Semiconductor Product Eng.
Infrared	Microwave & Storage Tubes
Anti-submarine Warfare	Communications Systems
Digital Computers	Inertial Guidance
Electro-optics	Field Engineering
Systems Design & Analysis	Circuit Design & Evaluation

Write in confidence to Mr. R. A. Martin
Hughes General Offices, Bldg. 6-E-11, Culver City, Calif.

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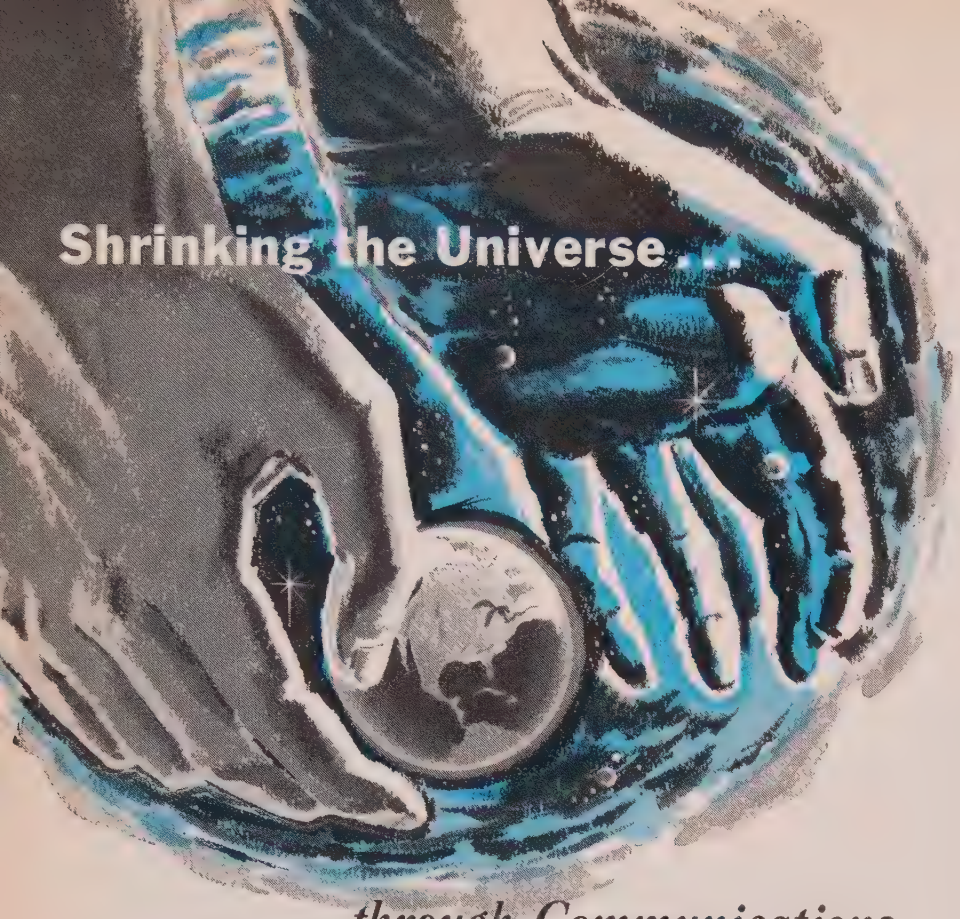
HUGHES

HUGHES AIRCRAFT COMPANY

Culver City, El Segundo, Fullerton, Malibu, Newport Beach, Oceanside, Santa Barbara and Los Angeles, California; Tucson, Arizona



through Communications

BY APPLYING TOMORROW'S CONCEPTS TODAY...
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today's dynamic R & D mission also includes creative opportunities in Navigation, Microwave Data Handling, ECM and Radar Systems

Man's vision of future exploration in the infinity of outer space is predicated on pertinent data that can be gathered by each space probe today. As the probe distances increase... our universe becomes smaller and communications are the vital link that proves success or failure of each venture into the unknown. It is imperative that new systems of communications be developed well in advance of other facets of space technology. At Amherst Laboratories new solutions are sought to communications problems even before they are encountered. Creative ideation is not compromised by conventionally accepted conclusions. Amherst Laboratories is currently anticipating tomorrow's electronics needs in areas of advanced Ground, Air and Space Communication.

PROFESSIONAL STAFF AND MANAGEMENT OPPORTUNITIES *are unlimited for Electronics Engineers, Physicists and Mathematicians with advanced degrees and creative desire.*

You are invited to direct inquiries in confidence to Mr. W. Serniuk, Director, Amherst Laboratories... or call NF 3-8315 for information. All inquiries will be acknowledged promptly.

SYLVANIA ELECTRONIC SYSTEMS
A Division of SYLVANIA
Subsidiary of GENERAL TELEPHONE & ELECTRONICS
1181 Wehrle Drive, Williamsville, New York

AMHERST LABORATORIES

Positions Open

(Continued from page 144A)

ELECTRONIC ENGINEER

Electronic Engineer with specialty in communication circuitry, in particular, IF filter design, ladder type networks and the construction of ceramic multistaged filters. Opportunity to head up entire division. Must have a man with a minimum of 5 years' experience. Salary depends on the individual. Box 2032.

ELECTRONIC CIRCUIT DESIGN ENGINEERS

Several years experience and graduate training desirable (but not required) for challenging circuit design problems. Ability to work in small, outstanding group on varied high caliber projects, e.g. computer techniques application, data accumulation and reduction, pulse amplifier and discriminator design, and systems analysis design. Phillips Petroleum Co., Atomic Energy Div., P.O. Box 2067-DO, Idaho Falls, Idaho.

TUBE ENGINEERS—PHYSICISTS—CHEMISTS

Astonishing new developments in microminiature tube design require research and development personnel non-Government work. Apply Kay Electric Co., Pine Brook, New Jersey. Tel. CA 6-4000.

ELECTRONICS ENGINEER

To be key member of small R & D group. Duties include design, development, testing experimental equipment and prototypes. Knowledge of automatic controls, high voltage rectifier equipment or gaseous discharges. Resume to be submitted to H. W. Buswell, Koppers Company, Inc. P.O. Box 298, Baltimore 3, Md.

EDITOR

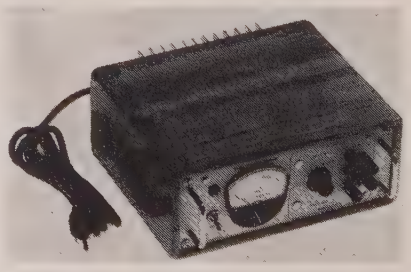
Scientific book publisher needs Editor to plan and develop books in radio engineering and electronics. Experience in publishing or science necessary; both desirable. Write in detail. State salary. Box 2034.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 28A)

Dual Power Supply



Two separate regulated outputs of 10 to 20 volts at 200 ma are provided in the Model 500 Dual Power Supply developed by **Burr-Brown Research Corp.,** Box 6444,

(Continued on page 154A)

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Operations research at ORO adds new dimensions to military tactics and strategy. Combining the diverse backgrounds of individuals representing more than thirty different disciplines, the Operations Research Office of the Johns Hopkins University employs modern scientific methodology on a mixed team basis in the solution of world-wide military problems in advanced and pioneering applications. The result: improved operational readiness and a better national military posture.

This calls for non-routine thinking about the broad spectrum of problems posed by the future demands of the modern Army. The areas of interest include strategy, war probabilities, tactics, logistics, intelligence, and air defense.

Scientists and engineers who wish to tackle operational problems of unusual scope and diversity are invited to join ORO. You will find the professional atmosphere of our organization conducive to creative thinking, and you will work with colleagues who have earned a national reputation as leaders in their fields.

Address your inquiry to
John G. Burke, Research Personnel Officer.

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OPERATIONS RESEARCH OFFICE / The Johns Hopkins University
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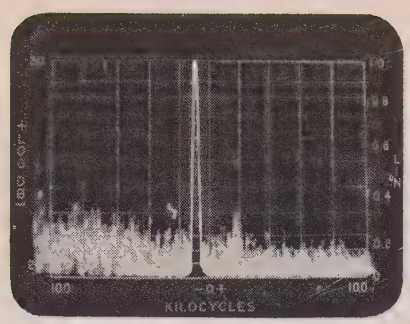

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The Operations Evaluation Group of the Massachusetts Institute of Technology is seeking men of proven ability in their chosen field who possess a firm grasp of the basic principles of mathematics or physics, the ability to interpret the results of basic research in their field, and the imagination to apply these research findings to the solution of present and long range problems which face the Navy. To these men we offer an opportunity to:

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- **Attack with scientists of similar backgrounds problems of critical importance in the area of National security.**
- **Enhance their professional standing in their chosen field by association with colleagues of the same discipline and by periodic academic study and research at M.I.T.**
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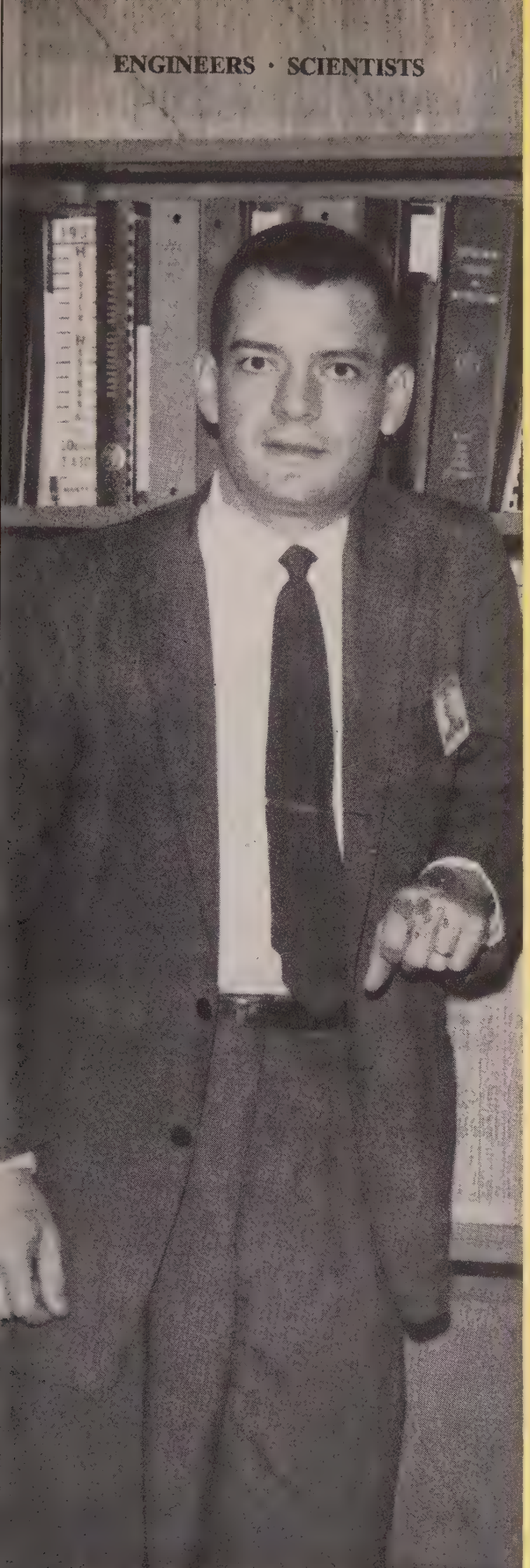

Dr. William C. King (Ph.D. Duke University) joined General Electric's Space Sciences Laboratory in 1956, after holding Research Associate appointments at Johns Hopkins and Duke. While with the Laboratory, Dr. King has published a number of papers, the most recent being in the area where he is now concentrating his interests: "Correlation of Telemetry Signal Attenuation Data with Transition from Free Molecular to Continuum Flow."

Communicare In Spatium
NO. 3 IN A SERIES

"THEORETICAL AND EXPERIMENTAL STUDIES OF COMMUNICATIONS IN THE PRESENCE OF HIGHLY IONIZED GASES ...a challenging interdisciplinary endeavor at the Missile & Space Vehicle Department"

SAYS Dr. William C. King, COMMUNICATIONS PHYSICIST AT GENERAL ELECTRIC'S SPACE SCIENCES LABORATORY

"Behind new emphasis at the MSVD Laboratory on the extension of knowledge in this little explored area of communications is the compelling need of space scientists to learn a great deal more about the physical phenomena encountered by a satellite or missile nose cone during re-entry.

"Every flight of a space vehicle should increase our store of information. This objective, however, can be met only if the results of measurements made in or near the vehicle can be delivered to the ground.

"Radio telemetry seeks to provide this link but during the crucial re-entry period the presence of highly ionized gases can make such communications difficult or impossible.

"The study—both theoretical and experimental—of this problem falls into not one but three scientific areas: electromagnetic theory, aerodynamics and physical chemistry. While the group I am leading is concerned principally with the first of these, it is necessary for us to work closely with those in allied areas.

"We are now moving into a period where new lines of investigation will soon be initiated, and are interested in augmenting our staff. A wide choice of problem areas is open to research physicists or electrical engineers, who wish to pursue new areas while working in an academic atmosphere, where the publication of original work is warmly encouraged."

Personal interview may be arranged with Laboratory managers by writing Mr. D. G. Curley, Div. 53 MK
MISSILE & SPACE VEHICLE DEPARTMENT

GENERAL ELECTRIC

3198 Chestnut Street, Philadelphia 4, Pa.

Opportunities now exist for Physicists and Electrical Engineers to engage in advanced studies requiring a background in investigations at doctoral level in one or more of these areas: analysis of interactions of electro magnetic waves with gases & plasmas; statistical properties of detecting signals in noise; coding systems; information theory; communications theory; propagation studies.

Other professional openings in: Aerophysics • Aerodynamics • Materials Studies • Structures and Dynamics • Plasma Physics • Magnetohydrodynamics

RESEARCH AND DEVELOPMENT AT LOCKHEED

...is setting the pace in space for years ahead

Lockheed Missiles and Space Division's progress transcends even that of an era marked by phenomenal scientific growth. To an important degree, the Division's research and development activities are considered to be the basis of its success.

As systems manager for the Navy POLARIS Missile and the Air Force AGENA Satellite in the DISCOVERER, MIDAS and SAMOS programs, the Division is engaged in extensive research in many diverse engineering and scientific fields. Some highlights of current research and development activities include: Operations research and preliminary design; nuclear and space physics; physical electronics; chemistry; materials; mathematics; engineering mechanics; electronic communications and instrumentation; and computer research and development.

Research is a concept which holds many different meanings to those concerned with science and technology. At Lockheed, a distinction is made between the *nature* of the work and its *objectives*. Consequently, such terms as basic research, applied research, systems or operations analysis, engineering and development are used. A given individual might find that his personal inclination often leads him quite naturally from one type of research to another. Recognition of this desire is reflected in the scope of work conducted in the Research Branch at Lockheed Missiles and Space Division. Principal research activities are: Pure and applied research; advanced design; engineering analysis; electronic prototype development; and machine computation.

Organization is determined by the *technical field* rather than by the *type of research*. For example, a structural dynamicist, as a member of the Structures Department, may, on one occasion, work on future space vehicle configurations, at another time be associated with current projects such as the POLARIS or Satellite programs, or he may be engaged in basic research at the research laboratory. In each case, the individual has the opportunity to maintain as much or as little contact as he wishes with others in his field of interest.

Important staff positions at Lockheed's Research and Development Branch in Palo Alto are available. Those scientists and engineers with experience related to the above areas are invited to write to: Research and Development Staff, Dept. 1-16, 962 West El Camino Real, Sunnyvale, California. U. S. citizenship or existing Department of Defense industrial security clearance is required.

Lockheed / **MISSILES AND SPACE DIVISION**

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CAPE CANAVERAL · FLORIDA · HAWAII

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Physicists
Optics Engineers:

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COMMERCIAL ELECTRONICS
for

- **ELECTROMAGNETIC FIELDS ENGINEER**

  Design of sweep circuits for conventional and experimental display devices. Engineers or Physicists should have a strong interest in practical application of higher math theory to electron ballistics and electron optics problems.

- **PHYSICIST**

  Must have broad experience and interest in optics and sound understanding of basic principles. Will develop Schlieren systems and explore new applications of physical optics. The difficulty of the problems involved requires a PhD degree plus at least 5 years experience.

- **OPTICS ENGINEER**

  Must be able to design and specify projection optics and have an understanding of Schlieren optics. Will develop projection system, including light source, collector, projection and Schlieren optics, front and rear projection screens. Will *not* be expected to design lenses. A strong background in optics including 5 years experience is required.

For full information,
please address your inquiry in confidence to
Mr. A. E. Constantine,
Dept. 53-MK, Bldg. 2.

TECHNICAL EMPLOYMENT OFFICE

GENERAL ELECTRIC

Electronics Park, Syracuse, New York

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 148A)

Tucson, Ariz. The unit features 0.2% regulation, current and voltage monitoring, electronic overload protection, and a thermal cut-out for high ambient temperature. Packaged in a "half-rack" module, 3½ H×9½ W×11″ D, the Model 500 may may be rack mounted or used in its cabinet as an individual instrument.

Price and availability—$365.00. Delivery is from stock to 60 days after receipt of order.

Bistable Amplifier

A new ultra-sensitive Bistable Amplifier with unusually high power output has been announced by **Norbatrol Electronics Corp.**, 356 Collins Ave., Pittsburgh 6, Pa. The new Bistable is a fast-response, completely-static device with far greater reliability and sensitivity than can be obtained with mechanical relays, which the bistable unit may be used to replace in many regulating and control applications. The new device is a combination magnetic amplifier-controlled rectifier unit using all-silicon solid state components.

(Continued on page 158A)

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B. S. or M. S. in Physics, semiconductor experience desirable but not absolutely necessary. Ph. D. in Physics with evidence of competency in Solid State physics, related thesis work or published articles.

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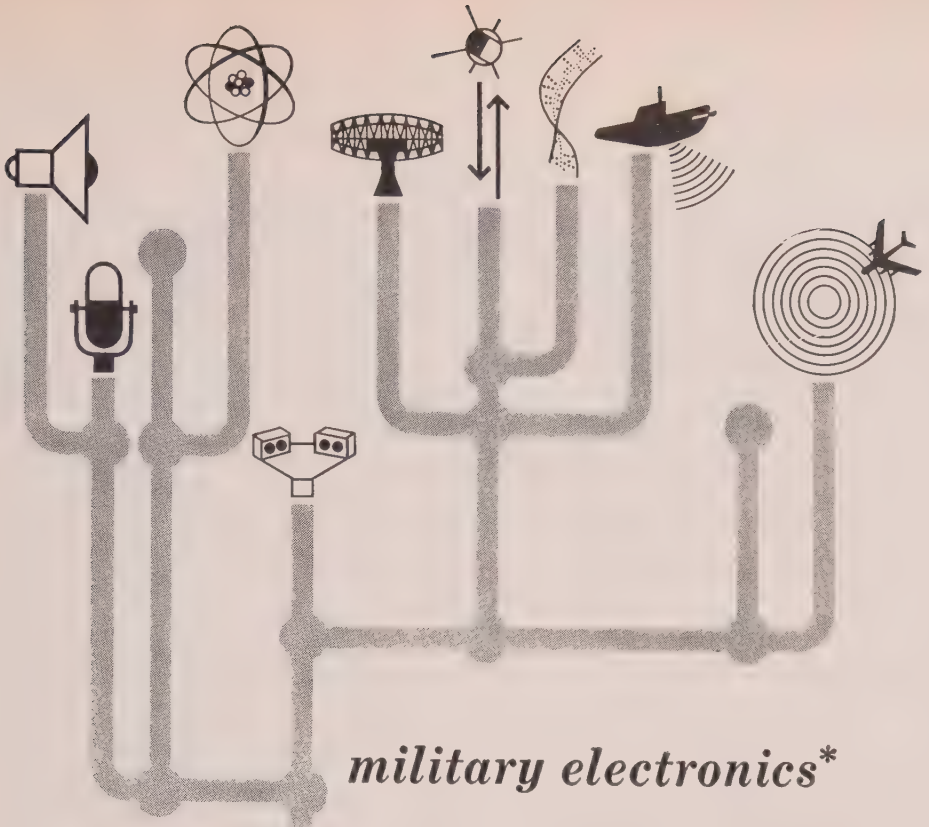

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DIVISION OF GENERAL DYNAMICS

New Professional Opportunities Now Open Across Our Entire Organization

The diversity and balance of our activities at Stromberg-Carlson – *in commercial and military fields* – creates both company strength and long term career stability for our professional personnel. *Growth* in both areas proceeds from a far-sighted policy of rising investment in research, to ensure the Company of an advantageous position as creator and supplier of continually more sophisticated systems and equipment. Specifically in the last five years, since Stromberg-Carlson became a division of General Dynamics, research budgets jumped 50%, product sales have risen dramatically, and our entire organization has expanded 60%. This year we are raising our research activity another 10%.

Engineers and scientists interested in the career-advantages of association with a broad-based, stable and healthily expanding company, while contributing professionally to challenging technical programs, are invited to inquire about new positions in their fields.

Please address Mr. Maurice Downey.

Areas in which Stromberg-Carlson is currently engaged include:

- **Nuclear Instrumentation**
- **Passive Reconnaissance**
- **Sonar Systems**
- **Automatic Test Equipment**
- **Stereo High Fidelity**
- **Single Side Band Communications**
- **Telecommunications**
- **Airport Traffic Control**
- **Avionic Systems**
- **Industrial Sound Systems**
- **Classified Investigations in ASW Field**
  (utilizing World's Largest Indoor Underwater Acoustic Test Facility inaugurated at Rochester, N. Y. headquarters in 1959)

STROMBERG-CARLSON

A DIVISION OF **GENERAL DYNAMICS**

1476 N. Goodman St., Rochester 3, New York

SCIENTISTS
ENGINEERS
CALIFORNIA

offers you and your family

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- Some of the nation's finest public schools for your CHILDREN
- World Famous Universities for ADVANCED STUDY
- MAJOR CULTURAL CENTERS

while living in such places as

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Fabulous Southern California
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companies pay interview, relocation and agency expenses

submit resume in confidence to:

PROFESSIONAL & TECHNICAL RECRUITING ASSOCIATES

(a division of the
Permanent Employment Agency)
825 San Antonio Rd.
Palo Alto, Calif.

GROW *with kollsman!*

AIR DATA INSTRUMENTS & SYSTEMS

Senior Project Engineers, EE & ME. For aircraft and missile instrumentation. 5 to 10 years' project experience in precision electromechanical devices, pressure transducers.

PRODUCT ENGINEERS

EEs & MEs. With product engineering experience on precision electromechanical systems.

AUTOMATIC ASTRO TRACKING SYSTEMS

Project Engineers, EE. For automatic astro tracking systems. Up to 5 years' related experience.

STAFF ENGINEERS & SPECIALISTS

a) Experience in the research and development of transistors in servo, digital and instrumentation application. Minimum 3 years' experience desired in transistor circuit design for military applications.
b) Experienced with IR to UV radiation properties and applications, noise theory and detectors.
c) Digital computers—logic or packaging experience.
d) Theoretical mechanics—inertial and trajectory studies.

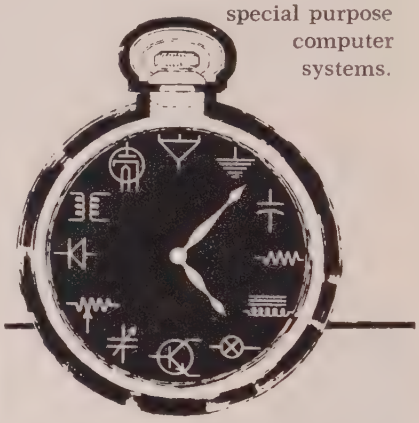
Please send resume to T. A. DeLuca.

kollsman INSTRUMENT CORPORATION
80-08 45th AVENUE, ELMHURST, NEW YORK

SUBSIDIARY OF STANDARD KOLLSMAN INDUSTRIES INC.

Electronic Creativity is a timeless quest

Boldly, significantly—intellectually curious scientists and engineers seek the most imaginative approaches in evolving new concepts and designs of technological electrocraft. At Melpar, the results have been such outstanding electronic achievements as the *first* integrated electromagnetic reconnaissance system ever developed, and *two* of the *largest* special purpose computer systems.



These examples of electronic creativity are typical at Melpar, where scientists and engineers are encouraged to apply new theories and fresh, perceptive concepts in the development of advanced electronic techniques.

For further information about provocative positions in the fields of space dynamics, including navigation, guidance, and simulation; pattern recognition techniques; advanced system design techniques, including extension of conventional machine organization concepts; and digital data handling systems—and for descriptive Melpar brochure—write to Frank J. Drummond, Manager, Professional Placement, 3335 Arlington Boulevard, Falls Church, Virginia, in historic Fairfax County, 10 miles from Washington, D. C.

An Arsenal of Technology
MELPAR INC
A Subsidiary of Westinghouse Air Brake Company

EET

Sandia Corporation's research in the field of piezoelectricity, ferroelectricity and ferromagnetism has opened a new field in "Explosive-to-Electric Transducers".

Sandia Corporation has developed high energy density transducers which convert explosive shock energy into megawatt peak electrical power.

Explosive transducers are typical of the provocative and rewarding fields at Sandia for men with ability and interest in working in the forefronts of research and development.

For those with a Ph.D. in physics or engineering Sandia offers an opportunity in a variety of fields including solid state physics, magnetohydrodynamics, materials research, explosives technology, pulse phenomena and radiation damage.

If you are interested in a career with Sandia Corporation write to either our Albuquerque or Livermore Laboratory for the illustrated brochure detailing the opportunities offered to these scientists and engineers.

Send Resume to Staff Employment Section 554

ALBUQUERQUE, NEW MEXICO
LIVERMORE, CALIFORNIA

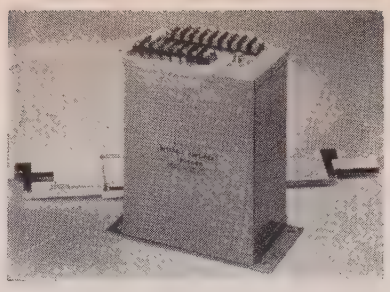
SANDIA CORPORATION

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 154A)

Responding to input signals of 0.05 microwatts, the amplifier produces a 24-volt dc output at currents up to 1 ampere. Response time is 22 milliseconds. The standard model has four control windings, permitting comparison, addition or subtraction of up to four different input signals.



In its simplest use, the new "static relay" acts as an ON-OFF device in a control or indicating system, amplifying a 0.05 microwatt signal from a thermistor, photocell, pressure transducer or other low-level sensing device.

In a regulating system the device functions like the relay voltage regulator in an automobile. The fast-response high sensitivity ON-OFF device provides optimum regulation in a proportional system.

The component is available with or without adjustable internal reference. Operating temperature range is −55°C to +85°C. The use of all-silicon semiconductor devices provides maximum protection against damage from overloads and exposure to high temperatures. Optional versions are available with extra control windings, output ratings up to 220 volts at 4 amperes, and a choice of ac or dc output.

Chopper Amplifier

The Chopper Amplifier Model 1801-0100 was specifically designed by **M. Ten Bosch, Inc.,** 80 Wheeler Ave., Pleasantville, N. Y., to convert a low level dc signal to a proportionate 400 cps signal at a much higher power level. This ac output signal can be conveniently used in the operation of a transistor amplifier such as MTB Model 1800-0300.

The Champ* requires a dc supply voltage of 28 volts at 3 ma with an ac reference voltage of 115 volts, 400 cps 3 ma. This unit has a nominal input impedance of 5000 ohms but a wide range of impedance values is available to suit requirements. The linear range is 0 to ±1 volt dc and the allowable range is is 0 to ±30 volts dc.

Output frequency is in the range of 380 cps to 420 cps and output voltage is 600 mv RMS maximum into a 10,000 ohm load. Variable gain is available by use of external resistance.

Size is $1\frac{1}{2}\times1\frac{11}{16}\times3\frac{15}{16}$″ and weight is 6 ounces. Unit meets military requirements within the temperature range of −55°C to +71°C.

*Trademark

Shift Registers Bulletin

Magnetics Research Co., 255 Grove St., White Plains, N. Y., manufacturer of digital magnetic devices, announces the availability of a four-page engineering bulletin on its Wide-Width Magnetic Shift Registers.

The bulletin includes a technical discussion of wide-width theory, as well as circuit diagrams explaining the functioning of its unique Ultra-Flex data-control circuit, which reduces the equipment re-

(Continued on page 163A)

GENERAL ELECTRIC ANNOUNCES IMMEDIATE OPENINGS FOR . . .

Transistor Development Engineers

Positions offer excellent opportunity for Engineers and Scientists who prefer a variety of technical responsibilities . . .

The Semiconductor Products Department of the General Electric Company has new openings for Transistor Device Development Engineers. These are permanent, career opportunities where your professional talents will be applied on semiconductor devices for the industrial, home-entertainment, and defense markets. And each offers unusual growth potential—in a business that is doubling every three years.

Qualifications:

All levels of experience in developing electronic components will be considered. However, you must be a technical graduate (B.S., M.S., E.E. or Physics). Citizenship is not required but permanent visa status is desirable.

Responsibilities:

You will become part of a team operation responsible for a variety of technical requirements —from semiconductor device inception to pilot-line stage including transistor design, process development and device evaluation. Even R&D work will be a part of your responsibilities.

This is an opportunity that should be very attractive to you if you are an Engineer or Scientist who is not interested in extreme specialization—especially, since the engineering function retains responsibility through the pilot-line stage. Your work will be in a range of areas and require a variety of technical contributions.

Many outstanding personal and professional benefits

- You work with General Electric's world-famous team of scientists and engineers now making history with a variety of new semiconductor devices.
- You can do graduate work under General Electric's Tuition Refund Program.
- You work in a diversified multi-market technology producing products for defense, industrial, and home entertainment fields.
- You and your family are protected by a comprehensive Company benefits program that is one of the most progressive in industry today.
- You live in a highly desirable geographic location with an outstanding school system, and served by a modern, major highway system.

FOR MORE INFORMATION contact in complete confidence:
Mr. J. H. McKeehan—Sect. 220-18
Semiconductor Products Department
General Electric Company
Electronics Park, Syracuse, N. Y.

AVAILABLE UPON REQUEST,

this brand new publication describes the wide variety of openings in General Electric semiconductors as well as interesting industry background, Company benefits, etc. Address your request to Mr. McKeehan.

GENERAL ELECTRIC

Inventiveness: indispensable ingredient of **S**pace **T**echnology **L**eadership

In the achievement of Space Technology Leadership, on-the-shelf hardware and the existing state-of-the-art are not always equal to the requirements of advanced missile and space systems. In such challenging situations Space Technology Laboratories responds with the full breadth of its resources ● In response to the need for time compression, STL inventiveness produced devices answering urgent requirements of advanced space programs conducted for the Air Force Ballistic Missile Division, National Aeronautics and Space Administration, and Advanced Research Projects Agency. Among these: Telebit, first digital computer to enter space; the first multi-million-mile space communications system of Pioneer V; a continuous-wave radio guidance system and lightweight autopilots for Able-series space vehicles; and a low-thrust multi-start space engine for maneuverable satellites ● On this foundation of inventiveness STL continues to broaden in scope, translating creative concept into accomplishment for Space Technology Leadership ● Outstanding scientists and engineers seeking such an environment are invited to investigate opportunities available at STL. Resumes and inquiries will receive meticulous attention.

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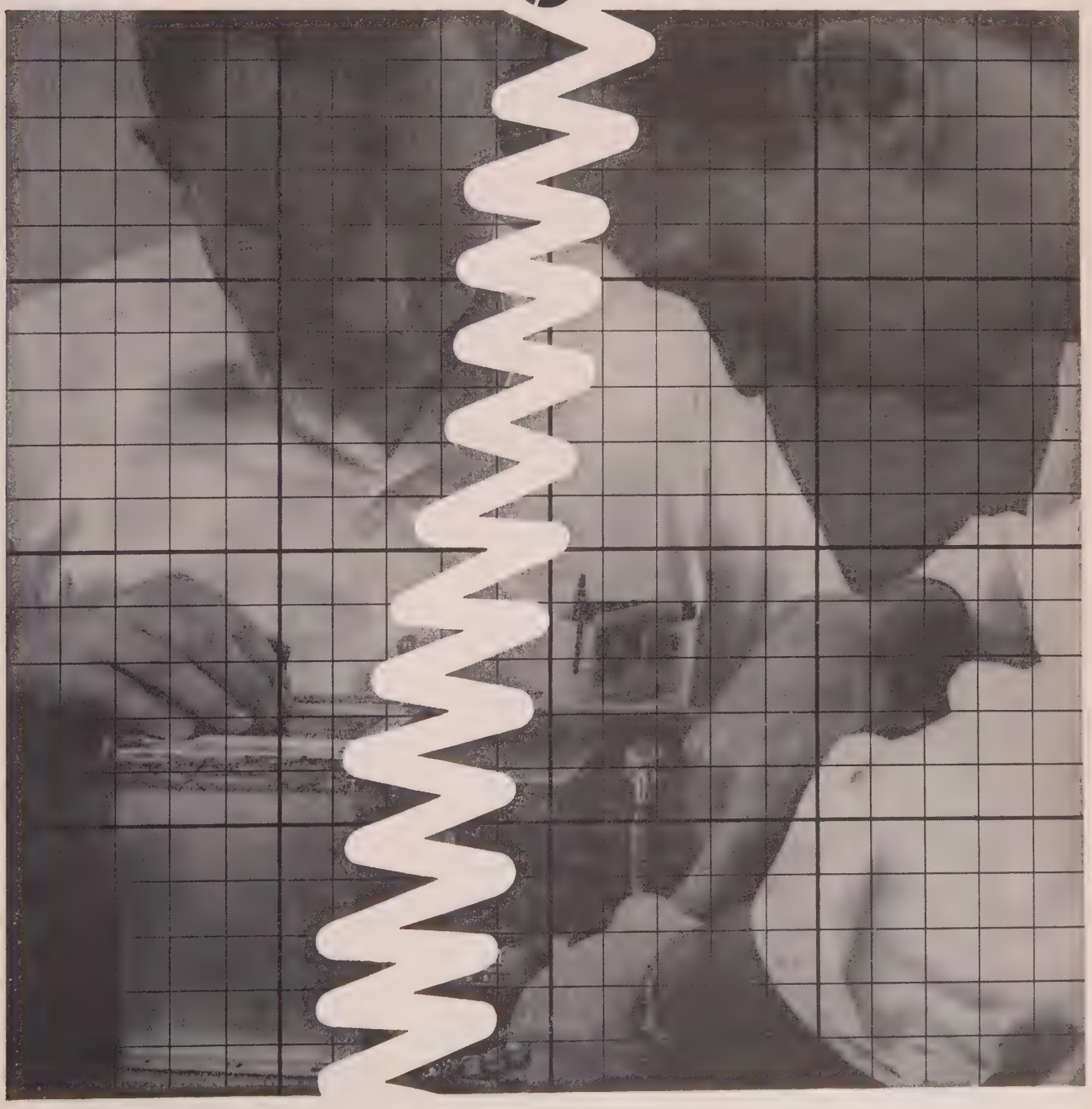

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We have embarked on an expansion program unprecedented in our 25-year history. It will take the form of accelerated research and development projects in the fields of:

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The man who will qualify will have a B.S. in electrical engineering (an advanced degree is desirable, but not essential), plus at least 3 years of applicable experience in development engineering. He will be capable of independent original thinking, and will instantly recognize the IMMEDIATE growth potential of this position. Because location is important too, he will welcome the opportunity to work in our modern plant in the lovely lake district of northern New Jersey.

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- *Continuing professional growth*
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There is an unmistakable degree of professional pride evident when a scientist or engineer says, "I am a part of the Naval Research Laboratory." To become a part of NRL requires a considerable degree of competence, of course. To learn how you may qualify for the many excellent job opportunities at NRL, please write to the Personnel Officer.

NRL

U. S. Naval Research Laboratory
WASHINGTON 25, D.C.

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Career Opportunities!

Your opportunity for professional advancement has never been greater than it is today at Remington Rand Univac.

Working in Univac's *Atmosphere of Achievement*, engineers and scientists are making significant advances in solid state development and computer reliability.

The following are a few of the positions immediately available:

- **DEVELOPMENT PROJECT MANAGER**
  This is an excellent opportunity for one who has had five or more years experience directly involved in the circuit design and hardware development of digital computing equipment.
- **MANUFACTURING PROJECT MANAGER**
  A very attractive position involving advanced manufacturing techniques used in deposition of thin magnetic films. Experience in magnetic circuitry as applied to memory and switch elements.

Other opportunities for scientific personnel with training in Engineering, Physics and Mathematics are available in the following categories:

- **TRANSISTOR CIRCUIT DESIGN**
- **SOLID STATE PHYSICS**
- **COMPUTER LOGICAL DESIGN**
- **COMPUTER PROGRAMMING**
- **BUSINESS DATA PROCESSING SYSTEMS**
- **MISSILE GUIDANCE AND CONTROL SYSTEMS**
- **ELECTRO-MECHANICAL DESIGN**
- **QUALITY CONTROL**
- **PRODUCTION ENGINEERING**
- **MINIATURE AIRBORNE ELECTRONIC PACKAGING**
- **ENGINEER WRITERS**
- **MILITARY PUBLICATIONS EDITORS**
- **COMPONENT STANDARDS AND SPECIFICATIONS**

R. K. PATTERSON, Dept. F-11

Remington Rand Univac

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2750 West Seventh St., St. Paul 16, Minn.

There are also immediate openings in all areas of digital computer development at our other laboratories. Inquiries should be addressed to:

F. E. NAGLE, Dept. F-11
REMINGTON RAND UNIVAC
1900 West Allegheny
Philadelphia 29, Pennsylvania

D. CLAVELOUX, Dept. F-11
REMINGTON RAND UNIVAC
Wilson Avenue
South Norwalk, Connecticut

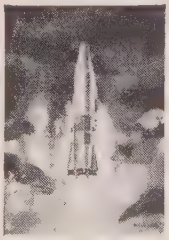
STATISTICAL COMMUNICATIONS THEORISTS

Convair Astronautics needs three men with either a Ph.D. or Sc.D. in statistical communication and information theory. These men are for senior level positions in our new Electronics Research Laboratories.

The task will be to perform research studies into advanced systems techniques for application to communication, tracking, data analysis, optical, and display systems such as required by satellites and advanced space missions.

A rigorous analytical approach using statistical communication theory, information theory, and sampled and quantized signals, power spectral analysis, and network theory is required. Write at once to Mr. R. B. Merwin, Industrial Relations Administrator-Engineering, Dept. 130-90,

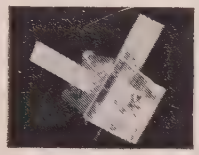
CONVAIR / ASTRONAUTICS



CONVAIR DIVISION OF
GENERAL DYNAMICS

5553 KEARNY VILLA RD., SAN DIEGO, CALIF.

An invitation for Physicists and Engineers to work on Research and Development at the M.I.T. Instrumentation Laboratory . . .



The Laboratory, with its staff of 1,000 employees, is primarily engaged in the conception and perfection of completely automatic control systems necessary for the flight and guidance of missiles and space vehicles. Many "firsts" in these fields have been developed at the Laboratory.

R and D opportunities exist in: • ANALYSIS OF SYSTEMS AND COMPONENTS • ASTRONAUTICS • HIGH PERFORMANCE SERVOMECHANISMS • POWER SUPPLIES & MAGNETIC AMPLIFIERS • DIGITAL & ANALOG COMPUTERS • ELECTRO-MECHANICAL COMPONENTS • TRANSISTOR CIRCUITRY & PULSE CIRCUITRY • RESEARCH, DESIGN & EVALUATION OF GYROSCOPIC INSTRUMENTS • COMPUTER PROGRAMMING & SIMULATOR STUDIES • LOGICAL DESIGN • AND IN MANY OTHER AREAS

CALL or WRITE:
DR. C. S. DRAPER, Director

INSTRUMENTATION LABORATORY
MASSACHUSETTS INSTITUTE OF TECHNOLOGY

68 Albany Street, Bldg. 11-A
Cambridge 39, Massachusetts
UNiversity 4-6900, Ext. 3544

- Graduate courses may be taken for credit while earning full pay
- U. S. Citizenship required

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offers career opportunities for challenging assignments in commercial products R & D to:

TRANSISTOR CIRCUIT DESIGN ENGINEERS

Digital Data Processing Electronic Engineers with practical experience in translating ideas to the hardware stage. Experience in systems, logical design, transistor circuitry, memory systems and input/output equipment is necessary.

ELECTRONIC ENGINEERS OR PHYSICISTS

Electronic Design Engineers interested in applying the unique capabilities of electronics to the complex of industries served by FMC. Informed communication with individuals in diverse fields is essential. Ideas for solutions to problems are carried through to the prototype stage. Academic training and work experience in general electronics required.

FMC Central Engineering Laboratories' major expansion program requires well-qualified engineers with a high degree of creative imagination to staff our new million dollar facilities in the San Francisco Bay Area. BS required and advanced degrees desirable for these responsible positions.

Interested? Send a resume of your background and professional experience to E. M. Card, Jr., FMC Central Engineering, 1105 Coleman Ave., San Jose, Calif., or telephone CYpress 4-8124 for an interview appointment.

fmc FOOD MACHINERY AND CHEMICAL CORPORATION

Putting Ideas to Work

Central Engineering Laboratories

FOOD MACHINERY AND CHEMICAL CORPORATION

006CE

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 158A)

quirements to one core per bit. It is believed by the firm that the bulletin contains the first published source of this material.

Typical characteristics of standard shift registers employing these principles are also given.

The bulletin is offered free to qualified engineers requesting it on their company letterhead. Address inquiries to Eugene Sands at the firm.

Wirewound Resistors Bulletin

Sprague Electric Co., North Adams, Mass., announces three series of glass-jacketed power resistors. The Series F units, comprising Types 750E, 751E, 752E, 753E, 754E, 755E, and 756E are inductively wound with either single- or multiple-layer windings, depending on the resistance value. Series NIF units, comprising Types 757E, 758E, 759E, 760E, 761E, 762E, and 763E, employ "Ayrton-Perry" windings to minimize the self-inductance. These are customarily referred to as "non-inductive" resistors.

(Continued on page 164A)

Primary Standards Measurement Engineers

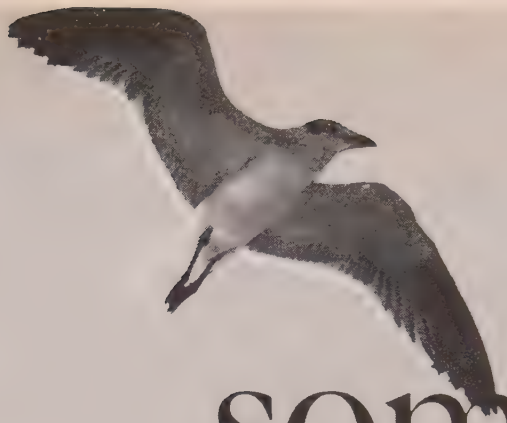
Boeing, major contractor on such advanced programs as the Dyna Soar boost-glide vehicle and the Minuteman ICBM, has a number of openings in the field of Primary Standards Design Analysis. Requirements are an advanced degree in ME, EE or Physics; or a BS degree plus experience in fields related to research and applied metrology.

Assignments are available in Research Design Analysis of Precision Measurement Instrumentation in:

ELECTRICAL		
Voltage	Phase	Magnetics
Current	Frequency	VSWR
Resistance	Q	Attenuation
Power	Impedance	Field Strength
PHYSICAL		
Acceleration	Flow	Mass
Velocity	Temperature	Dimensional
Acoustics	Infrared	Pressure
Vacuum	Optics	

Send your resume, today, to: Mr. W. B. Evans, Boeing Airplane Company, P. O. Box 3707 - PRI, Seattle 24, Washington.

BOEING



some men see only gulls...

but a man of imagination named Otto Lilienthal saw more, and conceived an idea of consequence to the future of the world. The sources of world-shaping ideas are myriad; the men capable of conceiving these ideas, few.

In this age, more than ever before, there is a need for men of ideas. Now, at the new Autonetics Research Center in Whittier, California, a group of men are working to apply advanced physical concepts to the electromechanical technology of tomorrow. More men with advanced degrees are needed for this important work. Men to do research in moletronics, micromagnetics, advanced materials, and cryogenics.

Perhaps you are a man with new ideas . . . a man of imagination . . . one who requires the stimulation of new challenges. If you are, we invite you to share in the work at the new Autonetics Research Center.

Write: I. R. Knutson, Autonetics Research Center, 12000 E. Washington Boulevard, Whittier, California.

Current Openings for Experienced Ph.D's:

Solid State Physicists • Physical Chemists • Electrochemists • Physical Metallurgists • High Vacuum Technologists.

Autonetics

A DIVISION OF NORTH AMERICAN AVIATION, INC.

the heavens will be your workshop

At Cornell Aeronautical Laboratory an important, growing program in Radio Physics affords *YOU* a number of challenging scientific opportunities. Evidenced by the recent opening of a new Radio Physics Laboratory, our expanding research effort already includes such programs as satellite-to-ground measurements of transmission path characteristics, investigations of lunar and auroral reflection-structure characteristics via bi-static uhf transmissions, deep-ionosphere probing with the free-world's most powerful peak-pulse radar, rigorous analytical attacks on problems in the field of gaseous electronics, and system and component development in the realm of high powered radar. Scientists and engineers well based in the analytical or experimental aspects of electromagnetic propagation, and professionally interested in the pursuit of these and other problems, will want to consider joining the Laboratory's technical staff.

CORNELL AERONAUTICAL LABORATORY, INC.
of Cornell University

J. P. Rentschler
CORNELL AERONAUTICAL LABORATORY, INC.
Buffalo 21, New York

Please send me a copy of "A Community of Science."

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City ..... Zone..... State.....

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RP

NEWS New Products

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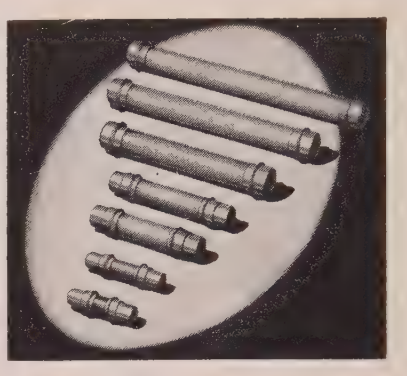
(Continued from page 163A)

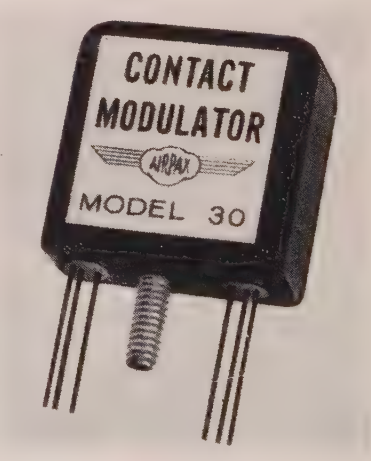


Super non-inductive resistors are known as Series SIF and comprise Types 764E, 765E, 766E, 767E, 769E, and 770E. These are obtained by an inter-leaved and "phased" multi-layer winding of Ceron wire. This results in extremely small phase angles at high frequencies. These resistors offer excellent performance characteristics for dummy rf loads, dummy antennas, line terminations, and rhombic-antenna phase resistors.

Sprague, on request, will provide you with complete application engineering service for optimum results in the use of these resistors in your particular circuit.

Micro-Midget Chopper

Weighing 9 grams, the diminutive, electro-mechanical Model 30 chopper developed by **Airpax Electronics, Inc., Cambridge Div.,** Cambridge, Md., introduces a new operating principle. Tailored for printed circuit use and employing ruggedized jewel bearings, the Model 30 is said to be the smallest low noise chopper in the world.



Characteristics—Drive: 6.3 volts at 60 cps; Dwell: 175°, average; Phase: $25 \pm 10°$; and Balance: within 15°.

Delivery is 2 to 6 weeks. Price on request. For further information, contact Walter D. Heisler at the firm.

(Continued on page 166A)

ENGINEERS • SCIENTISTS

Large-scale system engineering requires the anticipation of the need for design and integration of new equipments and subsystems to maintain a continually effective command and control structure.

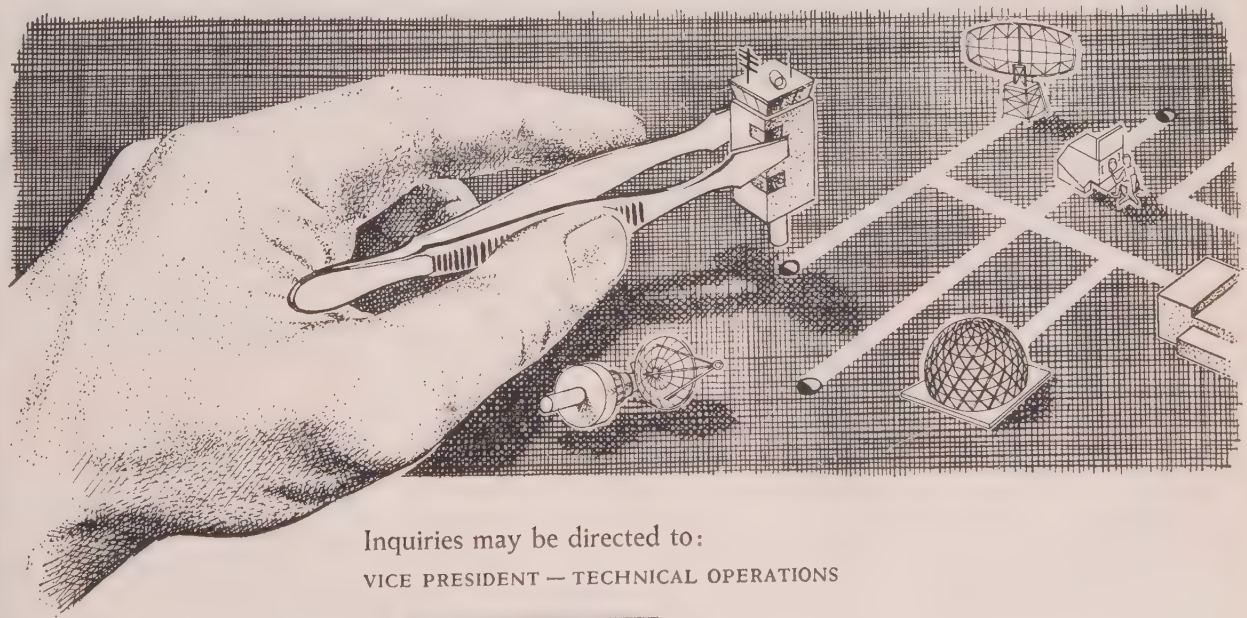
The elements of these systems — communication networks, computers, detection systems and weapons — must be continually evaluated, upgraded and integrated to assure total system compatibility. Effectiveness can only be maintained by the "open end" concept.

MITRE, a nonprofit system engineering organization, is engaged in the design, development and evaluation of electronic command and control systems.

Appointments to MITRE's Technical Staff are available to engineers and scientists with advanced degrees and interest in combining large-scale system engineering with electronic system development.

COMMAND AND CONTROL SYSTEMS...

EXPANDABLE TO 1970 AND BEYOND



Inquiries may be directed to:

VICE PRESIDENT — TECHNICAL OPERATIONS

THE MITRE CORPORATION

Post Office Box 208, 4-MM — Bedford, Massachusetts

ENGINEERS

Bausch & Lomb

A firmly established giant in the optical industry, this upstate New York corporation is rapidly expanding in electronic-mechanical systems work. Emphasis lies in commercial analytical and scientific instrumentation with a healthy balance of research and development contracts for military programs. Look into these professional opportunities with Bausch & Lomb Incorporated:

- **PHYSICISTS**

  Ph.D. or M.S. with experience and/or interest in the following areas: vacuum, solid state optics, infra-red, analytical instruments—to perform research in advance development.

- **ELECTRICAL ENGINEERS**

  B.S. with M.S. desirable. Three or more years experience in circuit design. Instrument experience valuable but not required. Must be capable of working without direct supervision.

- **MECHANICAL ENGINEERS**

  B.S. with M.S. desirable. Must have a strong background in machine design, small mechanisms and instruments. Must be capable of original product design and project responsibility.

  B.S. with M.S. desirable and a heavy background in physics and with design experience in the field of analytical or optical instruments.

- **INSTRUMENT ENGINEERS**

  Physicist with 2 or more years experience in optics and electronic instrument research and design.

  Mechanical engineer to design ophthalmic machinery, optical or analytical instruments.

- **OPTICAL ENGINEERS**

  Section Head — Optical Engineering M.S. or Ph.D. in physics or optics with broad background in optical, mechanical and electro-optical systems for military projects. To evaluate bid requests, supervise proposals, customer contacts, organize and direct projects.

  B.S. in Optics, M.E. or Physics and 2 or more years experience in optical engineering. Will provide the optical engineering design of a broad line of products.

- **LIVE IN ROCHESTER**

  Rochester is located in the heart of the upstate New York vacation country and is noted for its beautiful homes and gardens, fine schools and University, outstanding cultural advantages and high ratio of professional residents.

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  to forward your resume to H. A. Frye, Professional Employment.

Bausch & Lomb

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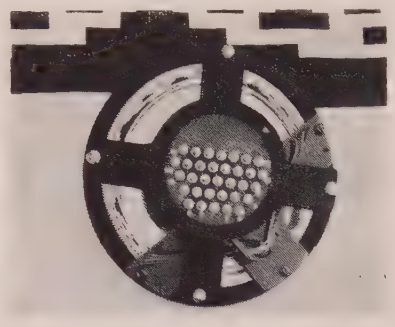
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 164A)

Shaft Position Encoders

United Aircraft Corporation's Norden Division, Wiley St., Milford, Conn., now is producing a shaft position encoder which provides 1024 counts per shaft rotation.



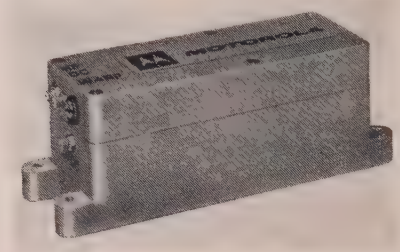
With an output in Gray Code and packaged in a standard BuOrd Size 33 synchro mount, the device supplements the manufacturers' existing line of Gray Code encoders which includes a Size 18 providing 256 counts and a Size 23 providing 512 counts per turn. All are one-half the diameter of encoders supplying an equivalent count in binary code.

The Gray Code configuration guarantees a non-ambiguous output signal without the need for V-Scan or other complicated circuitry at a consequent saving in size and cost.

Applications include those where minimum size encoders are of extreme importance and where computer circuitry for code translation is available.

High Stability Oscillator

A new high stability oscillator with a wide range of applications has been introduced by **Motorola, Inc.**, 4501 W. Augusta Blvd., Chicago 51, Ill. The hermetically sealed unit combines improved circuitry, physical design and crystal production techniques to assure stability of better than $\pm 1 \times 10^{-8}$ with minimum warm-up time and power consumption.



The oscillator can be used for single side band communications equipment, missile systems, navigation equipment, telemetering and instrumentation. It meets stringent military specifications for both

(Continued on page 168A)

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WRITE— Mr. J. W. PERRY
Professional Employment, Dept. 10

MARTIN
BALTIMORE 3, MD.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 166A)

airborne and ground equipment and operates at 3 mc (1 to 5 mc models available on special order).

Temperature control in the new device is of the proportional type using an ac bridge controlled by a thermistor. Rapid warm-up time and low power consumption are provided through the highly efficient oven which heats a chamber only large enough to hold the small crystal. Zener regulation provides stable operation essentially independent of power supply variations for the transistorized oscillator.

Magnetic Core Data Folders Available

A series of data folders has been announced by **Telemeter Magnetics, Inc.,** Components Div., P. O. Box 329, Culver City, Calif., providing complete specifications for a series of storage cores providing a wide range of characteristics. These data sheets describe both physical and magnetic characteristics with information concerning applications.

Copies are available from the manufacturer.

NEW, ADVANCED PROJECTS GROUP BEING ORGANIZED TO PURSUE RESEARCH AND DEVELOPMENT OF NON-CONVENTIONAL DISPLAY DEVICES

In the near future, a newly formed group with an unusual mission will be brought to full strength when close to 20 scientists and engineers with varied backgrounds join its staff.

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The group's output may find application in military, industrial or commercial fields. It will *not* be limited to any one . . . and may encompass all.

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- **Physical Optics**
- **Ceramic-Glass-Metal**
- **Vacuum Technology Development**
- **Tube Development**
- **Display System Analysis**
- **Circuit Analysis**
- **Synthesis & Application of Organic Materials to Electronic Components**
- **Geometrical Optics**

OPENINGS RELATED TO THE ABOVE ALSO EXIST IN OUR EXPANDING DESIGN ENGINEERING GROUP, TO WORK IN CLOSE LIAISON WITH THE ADVANCED PROJECTS OPERATION.

Write informally, or forward your resume in strict confidence to Mr. J. Wool, Dept. 53-MK. Your inquiry will receive prompt review and an early reply.

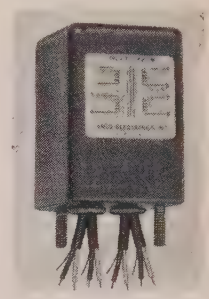
CATHODE RAY TUBE DEPARTMENT

GENERAL ELECTRIC

Electronics Park · Syracuse, New York

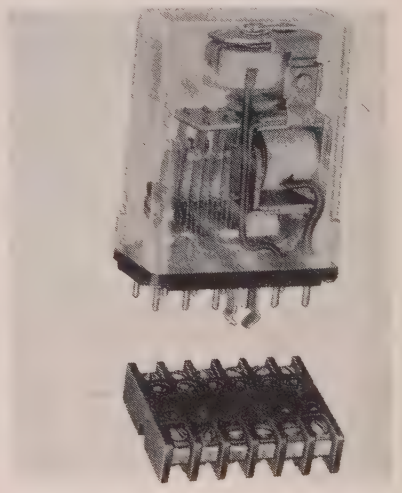
Chopper/Differential Instrument Transformers

A new line of chopper/differential instrument transformers incorporating several new design techniques is now available from **James Electronics Inc.,** 4050 N. Rockwell St., Chicago 18, Ill. Features include balanced windings (±0.05%) to the turn, high inductance (1200 Hys), high common mode rejection (120 db), hum bucking and magnetic shielding (135 db), low thermal construction, wide frequency response (30 cps — 8 kc) and non-microphonic epoxy impregnation of all models. The wide selection and features of these 4 shielded winding, 8 terminal transformers open new systems techniques to the instrument engineer. Complete technical brochure is available.



Memory Relay For Control

A new 255 Frame mechanical latch, electrical reset relay just introduced by **Struthers-Dunn, Inc.,** Pitman, N. J., is designed with clear plastic cover, 12-pin plug and matching heavy-duty socket for maximum economy, control simplification and minimum size for industrial control panels.

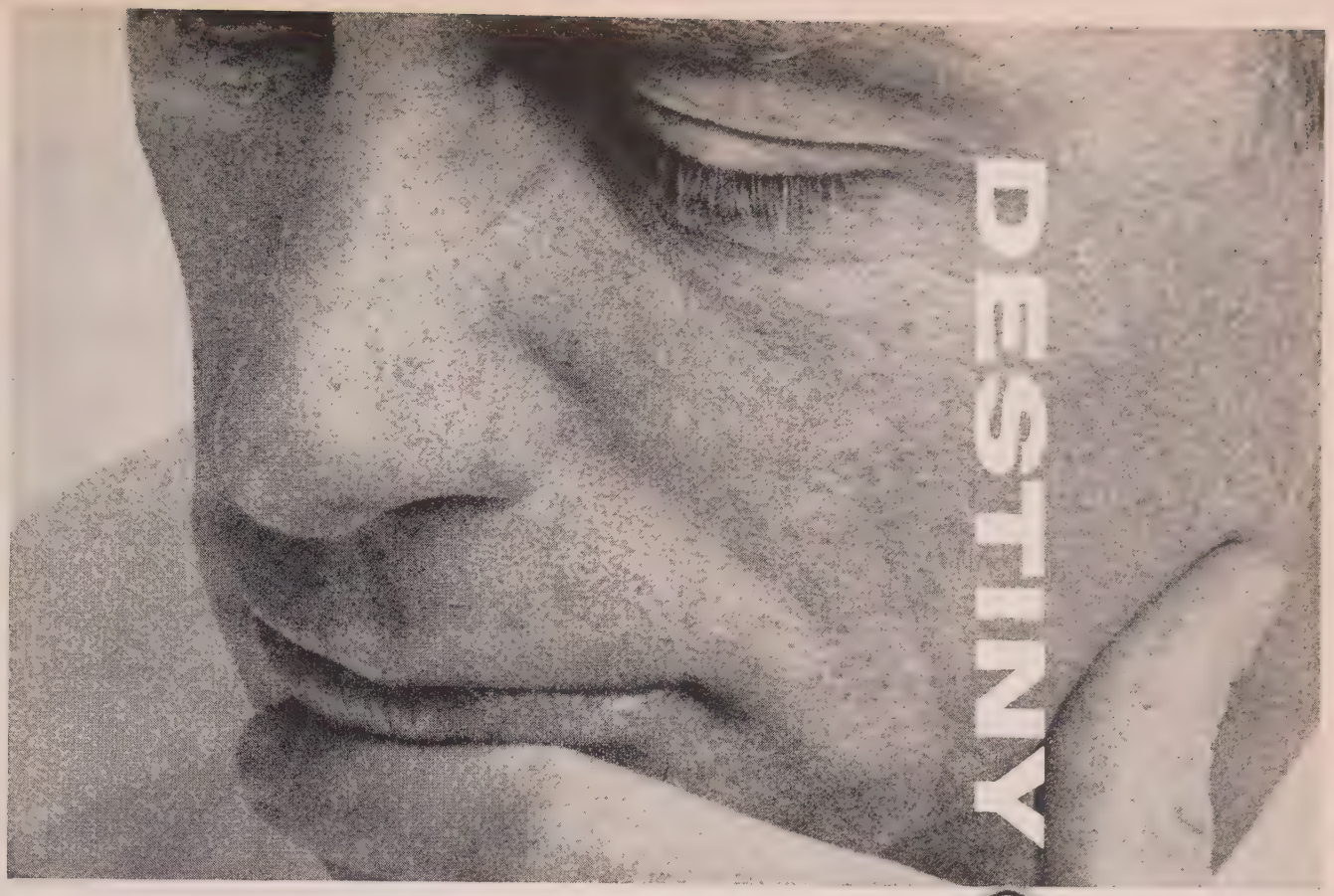


Similar to the "matching" Dunco 219 Frame "general purpose" and the 211 Frame "sequence" industrial control relays, the new "memory" unit consists of the widely accepted 219 frame plus a unique, sturdy latching frame. A rigid metal structure provides exact positioning of the two interfering armatures. This assures high contact reliability throughout a minimum life of 10 million operations.

Contacts are rated 10 amperes at 115 volts ac. All are operated by the same armature thus eliminating overlapping contact closures. Three types are available: with DPDT contacts; with TPDT contacts; and with two normally-open and two normally-closed contacts.

150 volt electrical spacings of $\frac{1}{4}''$ over the surface and $\frac{1}{8}''$ through air are used on

(Continued on page 170A)



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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

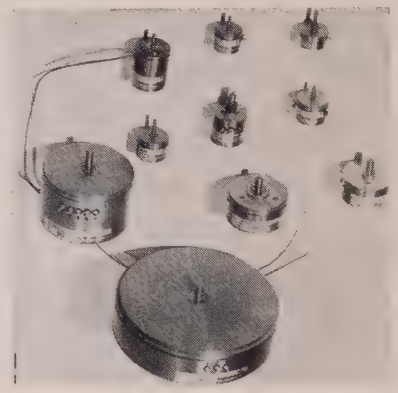
(Continued from page 168A)

all parts of the relay and socket. Standard dielectric test is 1500 volts ac.

The component is $3\frac{3}{4}$ high $\times 2\frac{21}{32}$ wide $\times \frac{27}{32}''$ deep

Full details are available in Dunco Data Bulletin 7255 available on request to the firm.

Potentiometer Bulletins

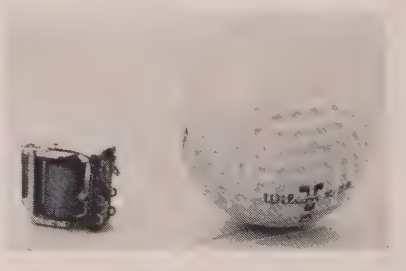


Accuracy, Inc., 4 Gordon St., Waltham 54, Mass., has published a series of data sheets which describe their new 2-Meg, Translatory and Protractatap potentiometers—all representing, among other characteristics, precise operation, high reliability and accuracy. Future technical data will be issued regularly.

This new potentiometer information is a "must" for design, test and production engineers and others seeking a new source for precision potentiometer design and manufacture.

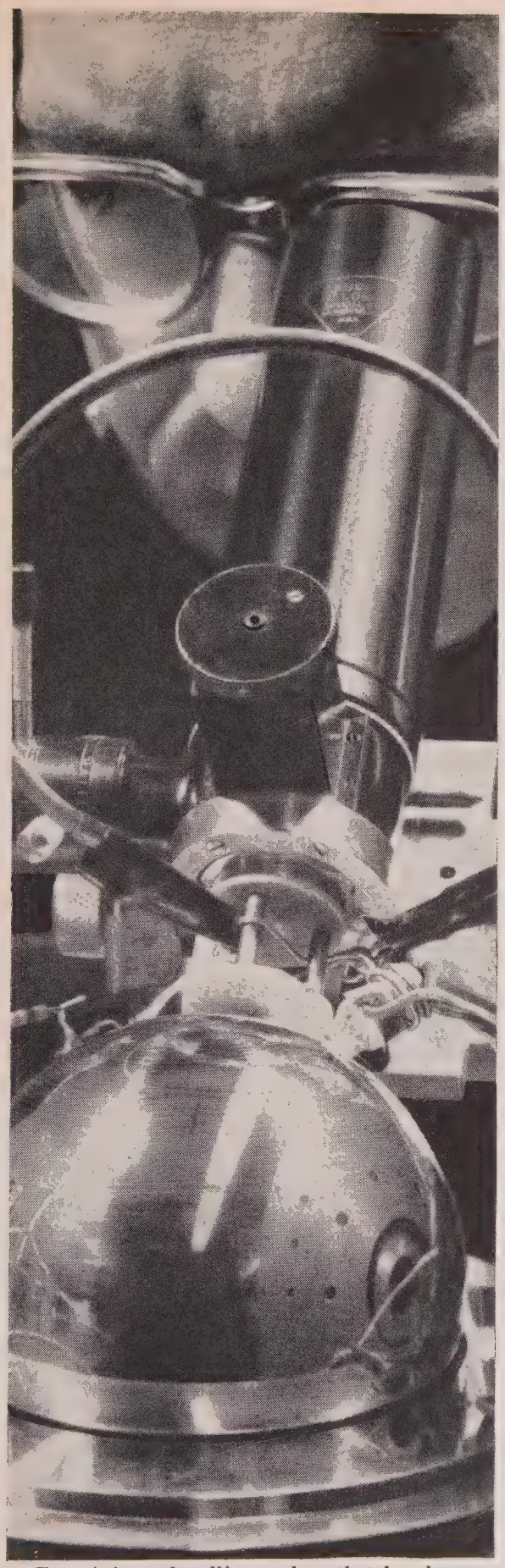
Subminiature DC Relay

High reliability, light weight, high sensitivity and small envelope are features of a new, low-cost subminiature SPDT dc relay manufactured by **W. S. Deans Co.,** 8512 E. Gardendale, Downey, Calif. The



relay has successfully passed a 6 cps, 508,500 cycle test with no visible sign of wear or pitting. A breakdown test of 115 volts ac at 250 ma was passed by contacts without arc suppression. Sensitivity: 35 milliwatts. Dimensions are: height, 0.687″; width, 0.406″; length, 0.812″. Other factors are: good contact wiping action;

(Continued on page 172A)



Examining a beryllium sphere that has been machined to tolerances finer than 15 millionths of an inch to form the rotor of an electrically suspended gyro.

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The continued expansion of Honeywell's Marine Systems Group has created a number of important new openings. This group is assigned the interesting and stimulating task of developing new electrically suspended gyros and submarine inertial systems using these advanced gyros. Openings include:

INERTIAL SYSTEMS ANALYST: Mathematician or engineer with strong background in vector analysis, operational calculus, matrix algebra and related techniques, to carry out analysis of inertial systems configurations including error evaluation.

DIGITAL SYSTEMS AND LOGIC DESIGNERS: BSEE or MSEE for design and development of electronic data processing equipment. Familiar with digital logic techniques at current state of the art; capable of organizing computing systems to perform various tasks including logical design and critical parameter specifications. Also engineers familiar with digital to analog data conversion.

ELECTRONIC ENGINEER: Electrical engineering degree plus experience in miniaturized semi-conductor electronics development. To design servo, pickoff, and other electronics for use with gyros and accelerometers.

ENGINEERING PHYSICIST: Physicist with practical and theoretical understanding of mechanics, magnetism and electricity to analyze and develop inertial sensors of novel and original design. Also experimental physicist with experience in the development of small, precise electro-optical devices.

MECHANICAL ENGINEER: BSME with 3 to 5 years' experience in extreme precision fabrication techniques to work on the alignment and stable mounting of precision optical-mechanical instruments.

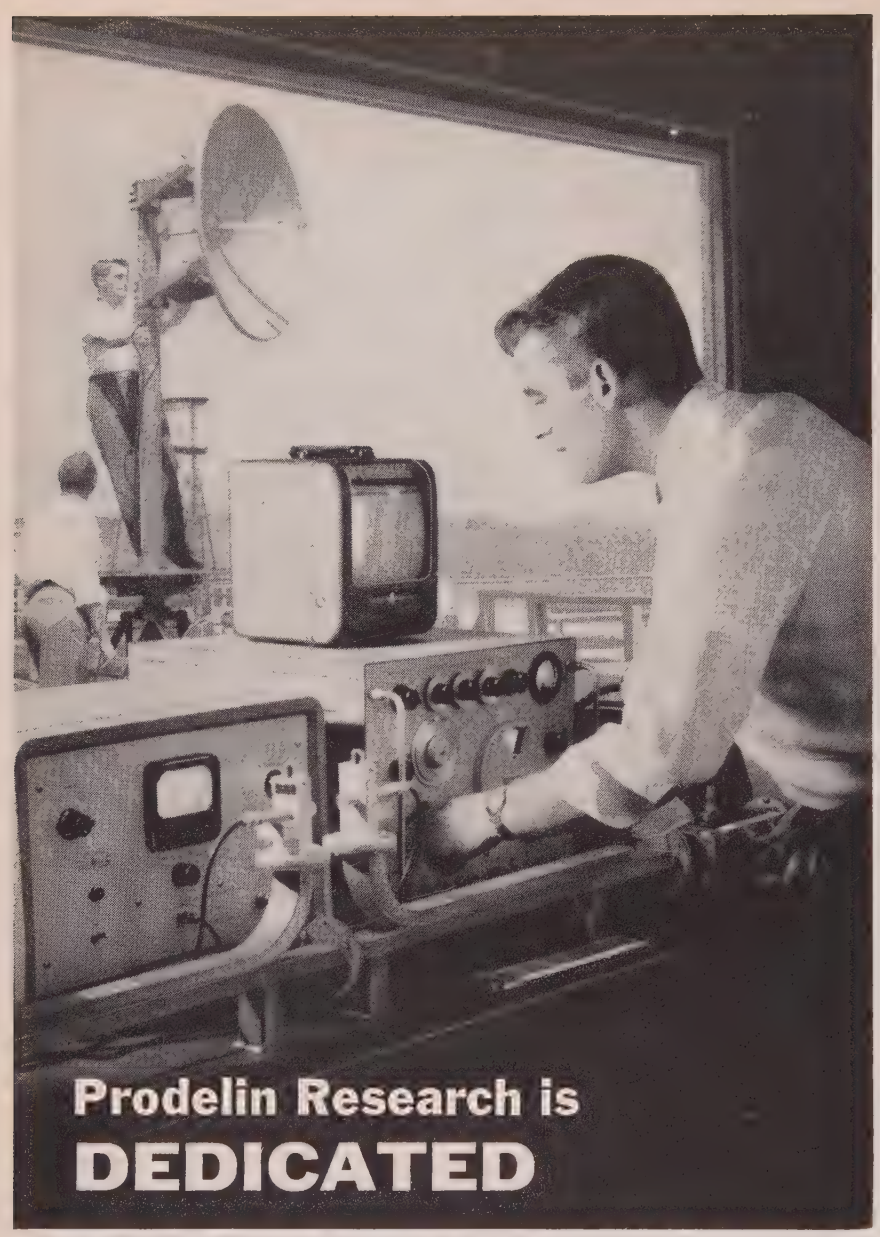
INERTIAL COMPONENTS ENGINEER: Familiar with inertial component specifications and available sources to conduct component evaluations for marine systems.

To discuss these or other openings, write Mr. James H. Burg, Technical Director, Aeronautical Division, 2600 Ridgeway Road, Minneapolis 40, Minn., Dept. 830.

Honeywell

H HONEYWELL *Military Products Group*

To explore professional opportunities in other Honeywell operations coast to coast, send your application in confidence to H. T. Eckstrom, Honeywell, Minneapolis 8, Minn.



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NEWS New Products

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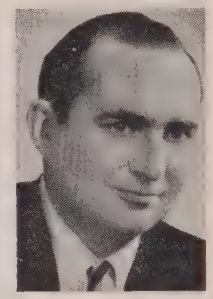
(Continued from page 170A)

weight 0.37 ounce; connection in single plane; single screw (2-56×¼) mounting. Units are available in 5000 ohm, 1250 ohm, 500 ohm, 100 ohm and 50 ohm coil resistances, or to user's specifications. For additional information on specifications, prices and delivery, write to the firm.

Loral Elects Garrett V.P.

Bernard R. Garrett has been elected vice president, engineering and research of **Loral Electronics Corp.**, 825 Bronx River Ave., New York, N. Y., it was announced by Leon Alpert, president and chairman of the board.



Garrett joined Loral in 1957 as assistant chief engineer and was named chief engineer in 1959. He holds a BSEE degree from New York University.

Loral is a major supplier of electronic systems and equipment for the Department of Defense. Through subsidiary firms it is a producer of electrical and electronic components for industrial and commercial applications.

Delcamp Appointed by FXR, Inc.

Raymond W. Delcamp, a physicist with nearly 20 years experience in government research, has joined **FXR, Inc.**, 25–26 50th St., Woodside 77, N. Y., manufacturer of precision microwave and associated equipment, as head of the microwave test equipment sales, William Marshall, vice president for sales, announced. Delcamp comes to FXR after 19 years as a civilian physicist with the U. S. Army Signal Corps research and development laboratory at Fort Monmouth, N. J.

For the past 10 years, he was chief of the laboratory's calibration equipment section, passing on all new equipment purchased. He was also responsible for maintenance and calibration of about 10,000 test equipment components. One of his major projects before leaving the Signal Corps was planning the establishment of a calibration lab covering an extraordinarily wide range from direct current to 40 kmc.

(Continued on page 174A)

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AN ELECTRONIC WHEATSTONE BRIDGE WITH DIGITAL ACCURACY (0.05%) For Only $795.00

- **Exceptionally wide measuring range — 1 ohm to 1,000 meg.**
- **"Calibrated unbalance" for quick reading of similar resistances in production testing.**
- **Compact rugged 12" x 8" x 9" case houses complete bridge, burn-out-proof electronic null detector and power source.**

As an accurate electrical resistance measuring device the 113-year-old Wheatstone Bridge has never been surpassed. Our best-equipped laboratories still use it for the ultimate in measuring accuracy. Their only concession to modern engineering has been the addition of a separate electronic null detector, strangely contrasting with the "old fashioned" look of the bridge it serves.

Millivac's Type MV-276A Electronic Wheatstone Bridge possesses all the commendable features of Sir Charles Wheatstone's original model. Moreover, it has a much wider measuring range — 1 ohm to 1,000 meg — made possible by the greater sensitivity of its built-in electronic null detector, unfortunately not available in Sir Charles' day. We invite your inquiry.

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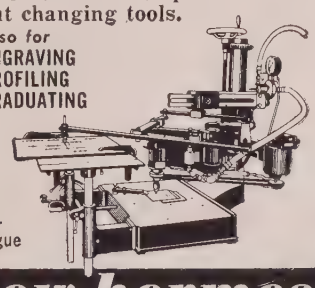

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NEWS New Products

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(Continued from page 172A)

Decade Scaler

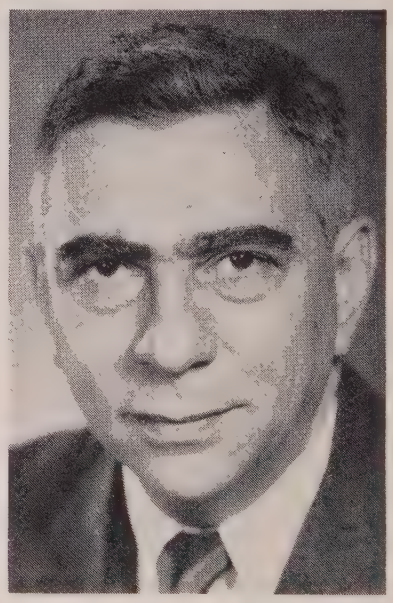
The Hamner N-240, 3½" decade scaler designed by **Hamner Electronics Co., Inc.**, P.O. Box 531, Princeton, N. J., primarily for use in analytical scintillation and proportional counting systems, where reliability, versatility and high speed of operation are essential.

The unit employs 4 plug-in decade strips and a 4-digit electro-mechanical register. Fast electronic stop-start circuitry is incorporated so that the scaler may be used with modern electronic timers. Fast gating ensures that scaler and timer will stop and start simultaneously. The N-240 can also be used with mechanical timers. Output connectors are provided for direct connection to either Hamner's N-803 or N-850 timers. Dimensions are 3½"×19"×13½". The scaler is operable in manual, preset time or preset count modes with an internal 60 cps calibration signal provided. Pulse pair resolution is 1 μs with a continuous repetition rate of 250,000 cps being possible.

The N-240 is available immediately. For detailed specifications contact the firm.

Piore Named V.P. by IBM

Dr. Emanuel R. Piore has been elected vice president for research and engineering, **International Business Machines Corp.**, 590 Madison Ave., New York 22, N. Y. Dr. Piore will continue as IBM director of research and in his new position will exercise staff supervision over the company's research and engineering activities.



Dr. Piore joined IBM in 1956. He previously was vice president in charge of research for the Avco Manufacturing Co. From 1946 to 1955, Dr. Piore was with the Office of Naval Research, becoming ONR's chief scientist. He is a member of President Eisenhower's Science Advisory Committee.

Plug-In Sweep Generator

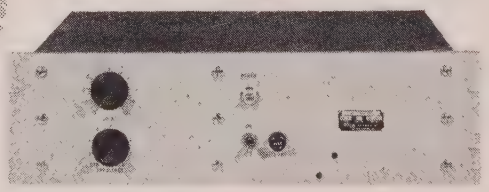
A sweep generator unit that attaches directly to and converts the majority of laboratory oscilloscopes has just been announced by **Teltronics, Inc.**, 277 Main St., Nashua, N. H. This instrument, useful for alignment of IF stages and wide-band amplifiers, requires no power cables or batteries since necessary energy is obtained from the oscilloscope. With no mechanical moving parts this equipment exhibits a high degree of reliability and requires a minimum of maintenance.

Features available include independent center frequency and sweep width controls, complete shielding, marker input provision, synchronization and blanking (through the oscilloscope), variable sweep

(Continued on page 177A)

IMPORTANT

TO BETTER COMMUNICATIONS



RIXON'S SERIES OF DELAY EQUALIZERS for continuously variable compensation of delay distortion. Data transmission requires faithful reproduction of pulses and is seriously affected by "delay distortion" introduced by wireline and other narrow-band networks. Rixon's EN-766 series of continuously variable Delay Equalizers compensates for delay distortion by passing phase distorted signals through all-pass networks having phase versus frequency characteristics complementary to the communications path. These compensating characteristics can be continuously varied until an oscilloscope presentation of the recovered signal agrees with its configuration prior to transmission. This insures rapid delay compensation by simple adjustment. Compensating characteristics centered at ten frequencies from 0.5 to 2.5 kc are available. A family of six additional characteristics equalizes pre-compensated circuits.

Rixon's thirty different models, along with special designs available will equalize most wireline delay characteristic.

Write for Engineering Bulletin

RIXON
ELECTRONICS, INC.
2414 Reedie Drive
LOckwood 5-4578
Silver Spring, Maryland

SPECIFICATIONS

Size / Width—19 in. rack standard; Height—3⅜ in.; Depth—7⅝ in.

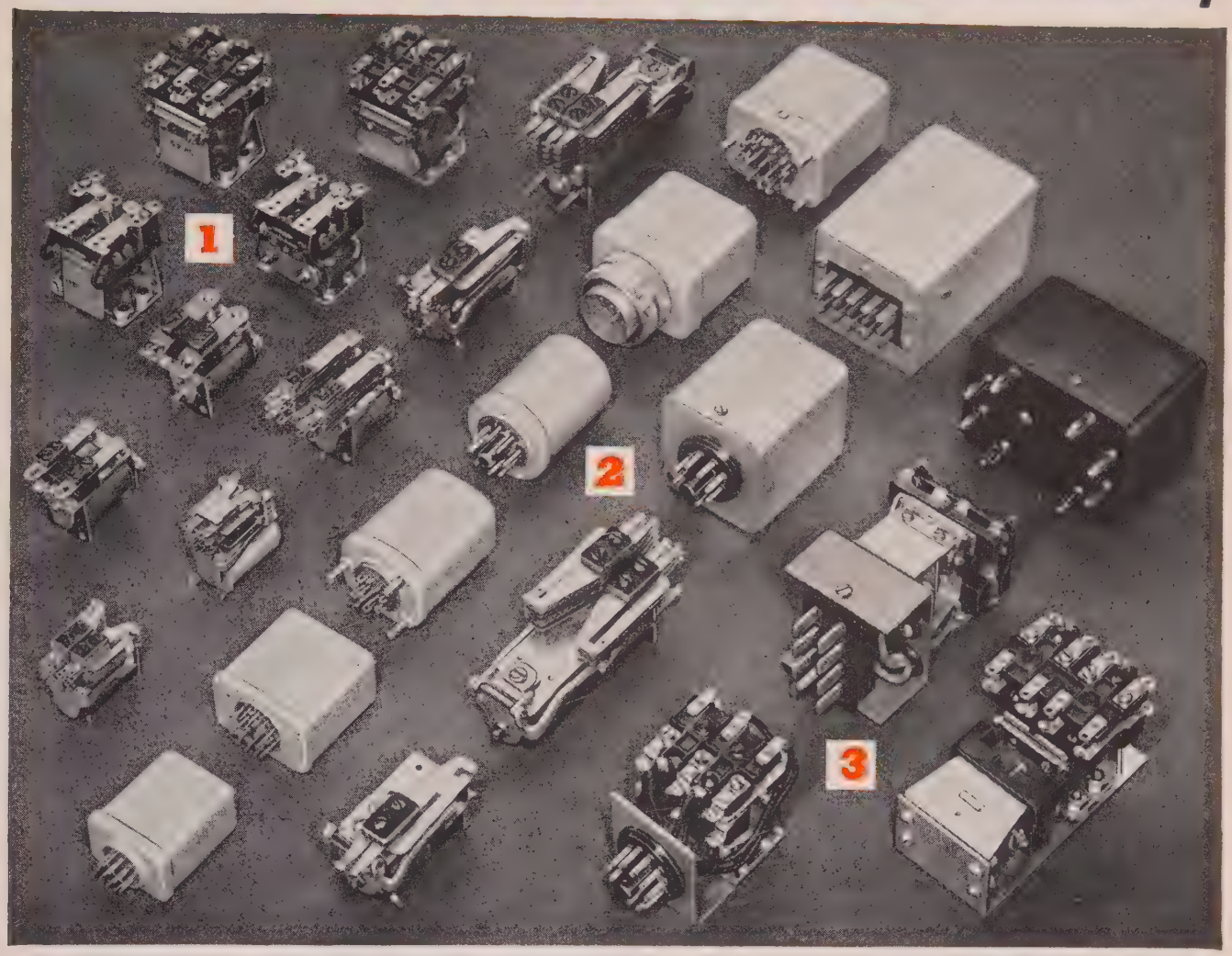
Delay Equalization/Effective over ranges from 0 to 3.0 kc with frequency of max. delay adjustable from 0.5 to 2.5 kc and delay range continuously adjustable from 0.3 to 3.5 ms.

Primary Input Voltage/100-130 V AC at 45-65 cps; 1152 AC at 60 cps nominal.

Primary Input Power/Less than 5 watts.

Input Level/+6 db max.
Input Impedence/600 ohms
Output Level/+6 db max.
Gain/0 db

specify OHMITE Relays for Reliability



1. Eleven Standard Models 2. Typical Relay Enclosures 3. Relays with Special Construction

EXCLUSIVE! Contact Combinations on New Ohmite Relays Are MOLDED*

Permanence and ease of adjustment of the individual contact springs are the result of a revolutionary, new innovation found in two new Ohmite Relays—Models TT and TS. This innovation is the unique "Molded Module"* contact spring construction. The "module" is a standard, single-pole, double-throw spring combination molded into a single compact assembly. As many as six modules can be incorporated into a relay. *(*Pat. Applied For*)

QUALITY CONSTRUCTION — All Ohmite relays embody the same meticulous engineering, strict quality control, and generous use of high quality materials which have made Ohmite components the standard of the industry. Parts are plated where necessary for corrosion resistance. Springs are of nickel-silver or phosphor-bronze. Contacts are fine silver. Special contact materials, such as silver, tungsten, palladium, or gold alloy, can be supplied. Protection against humidity and moisture is paramount and is accomplished in layer-wound coils, through complete sealing with cellulose-acetate. Relays are available in a wide range of coil operating voltages and contact combinations in both AC and DC types.

65 TYPES IN FOUR STOCK MODELS — For fast service, four models in the Ohmite relay line are carried in stock in 65 types at the factory, and by Ohmite Distributors from coast to coast.

HERMETICALLY SEALED AND DUST-TIGHT RELAYS —You can specify many of the basic Ohmite relays in nonremovable, hermetically sealed enclosures for applications requiring complete relay protection. These high-quality relays are sealed in seamless steel enclosures which are exhausted and filled with dry, inert gas under control of Ohmite engineers. Ohmite hermetically sealed relays are available with either plug-in or solder terminals. Relays are also made with nonremovable dust-tight covers and removable dust covers.

RELAYS WITH SPECIAL CONSTRUCTION —Ohmite relays are available with special terminals or special construction, such as relays with push-on or screw terminals, relays with binding-post terminals. Where quantities warrant, Ohmite will manufacture relays made to your specifications. Ohmite can furnish not only special terminals, special contact combinations, contact materials, and coils but also special enclosures, connectors, impregnation, or frames. Ohmite relays can be engineered to meet your special pull-in, drop-out, or time-delay requirements.

For your special or unusual relay applications, let Ohmite's experienced engineers help you work out the best solution.

Write on company letterhead for Catalog and Engineering Manual 58.

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RHEOSTATS · TANTALUM CAPACITORS
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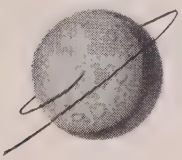
DIODES

COLLINS
ALPHA
CAPTURE AN 'ECHO' • TO BEGIN A NEW ERA IN RADIO

COLLINS, ALPHA CAPTURE AN 'ECHO' TO BEGIN A NEW ERA IN RADIO COMMUNICATION

The first two-way voice communication using the Echo I satellite was made by Collins Radio Company in Iowa and its systems subsidiary, Alpha Corporation of Texas. Collins also conducted signal measurements, relayed teletype messages and even a news photo. Participation in NASA's Echo project is representative of Collins' space research. Other studies involve upper atmosphere radiation, and systems evaluation associated with space vehicle programs, including a lunar "soft" landing. Collins also designed the first radio sextant and provided communication systems for Project Mercury and the X-15 rocket ship.

Challenging opportunities for engineers and scientists at all locations.



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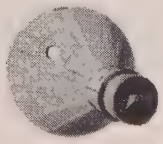

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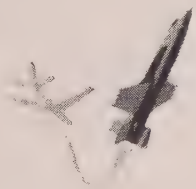

X-15 Electronics

COLLINS RADIO COMPANY
ALPHA CORPORATION

COLLINS

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NEWS New Products

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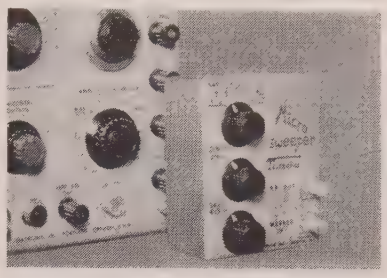


rate and 10^4 short term stability in center frequency. When the unit is used with a delayed sweep scope it is possible to measure transient response of circuits and filters. Also, the unit becomes a useful signal generator by setting the sweep width to zero.

For further information write to the firm.

IF Oscillator

A new IF Oscillator, known as the Model 139 IF Test Oscillator and designed for servicing and accurately aligning almost all radio receivers, including airborne and mobile communications equipment, has been announced as now in production by **Measurements, a Division of McGraw-Edison Co.,** Boonton, N. J.

The new precision instrument, a self-contained seven pound package, covers the commonly used high IF's from 3 through 20 mc in four continuous, individually calibrated bands, and also the industry standard low IF's in two independent crystal controlled frequencies, usually 455 and 290 kc, although other special frequencies are available.

The instrument can be used with or without a signal generator. For mobile radio use, it was designed as a companion piece to the Measurements Model 560-FM Signal Generator, a standard instrument for two-way communications alignment. Its features are said to include wide output voltage ranges, individually calibrated scales, and exceptional frequency accuracy and stability.

The Model 139 is priced at $165.00 with 455 kc crystal installed, second crystal optional.

(Continued on page 178A)

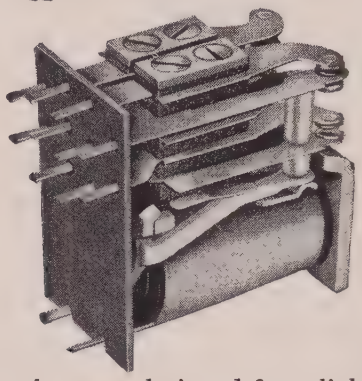
TEN STEP
5 megacycle stepping rates
PROGRAMMED
20 millimicrosecond pulse widths
CURRENT
2 ampere pulse amplitudes
PULSE
10 millimicrosecond rise time
GENERATOR

Model 1200 Programmed Millimicrosecond Current Pulse Generator for research and development of very high speed magnetic materials, solid state devices and computer circuits. *Write for Bulletin 60-B.*

rese engineering, inc.
A and COURTLAND STS., PHILADELPHIA 20
MAGNETIC CORE TESTERS • HIGH SPEED MEMORIES • LOGIC CIRCUIT PLUG-INS

Now...

A telephone type relay you can plug in and depend on!



Style 5400A

Style 5400A relays are designed for reliable switching in commercial and military printed circuit applications. They are available with contact combinations up to 4 form C and with a wide variety of contact materials for dependable switching of contact loads up to 5 amps.

An insulating board is provided to maintain printed circuit terminal alignment and to protect the relay from mechanical damage. Four tapped holes can be provided in the relay frame for additional mounting support if desired. All relays meet applicable portions of specification MIL-R-5757C.

General Characteristics:

Maximum Coil Resistance: *11,000 ohms*
Sensitivity:
Style 5409A (DPDT) 0.5 watts at pull-in.
Style 5424A (4PDT) 0.8 watts at pull-in.
Contact Combinations: *To 4PDT*
Military Specifications: *MIL-R-5757C*
Dielectric Strength:
1000 VRMS coil and contacts
Operate Time: *15 Milliseconds Maximum*
Release Time: *10 Milliseconds Maximum*
Temperature Range: *Standard construction to 85°C.; special to 125°C.*
Weight: *Approximately 1.7 oz.*

PRICE ELECTRIC RELAYS ARE QUALITY-CONTROLLED

For Additional Information, contact:

PRICE ELECTRIC CORPORATION

300 E. Church Street • Frederick, Maryland
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NEWS New Products

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(Continued from page 177A)

Sensitive DC Meter

Boonton Electronics Corporation, 738 Speedwell Ave., Morris Plains, N. J., has added the Model 95A Sensitive dc Meter to its line of dc measuring instruments. It features 17 voltage ranges and 25 current ranges. The full scale voltage ranges are from 10 μv to 1000 volts. The full scale current ranges are from 1 $\mu\mu$a to 1 ampere. The input resistance on the voltage ranges is 10 meghoms on all ranges. Accuracy is 3% of full scale on all but the most sensitive voltage and current ranges. On these ranges the accuracy is 4%.

A unique range switching and range indication system has been incorporated. One dial displays the full scale digits of the range selected in a window while a second dial displays the unit of measure in an adjoining window. This greatly simplifies the selection and recognition of any one of the 42 different measuring ranges. The measured quantity is indicated on a large zero center meter calibrated in the 1–3–10 series.

The dc output is available at the front panel. Separate controls for output level and zero reference, independent of the meter indication, give maximum flexibility in the use of external meters, recorders on control devices. Available from stock in September, 1960. Price $495.

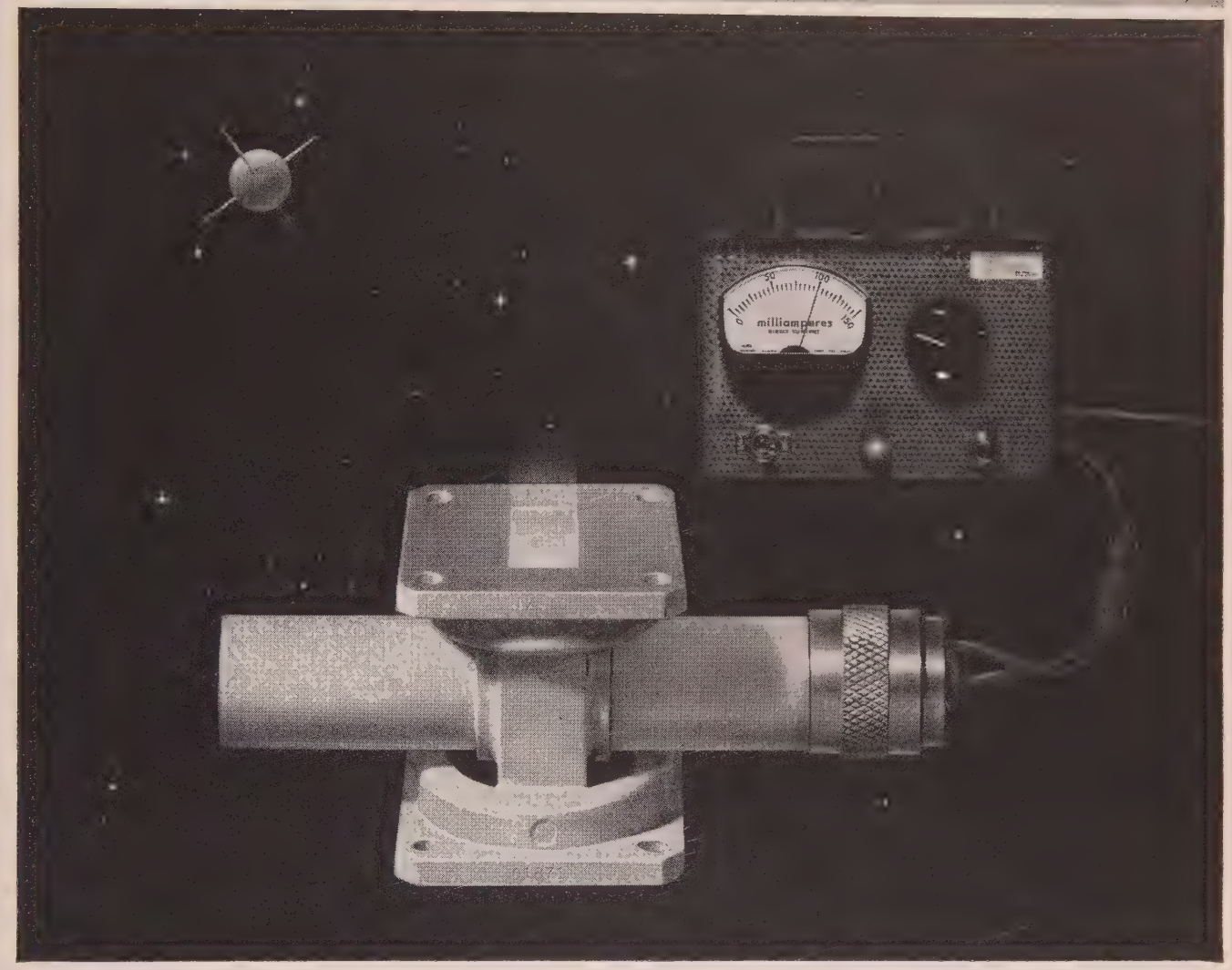
Diode Catalog

An eight-page, two color brochure/catalog No. SL-201/1 introducing more than 200 Planar diodes is available from **Fairchild Semiconductor Corp.,** 545 Whisman Rd., Mountain View, Calif. Listings include 115 standard 1N series and 103 FD series diodes with essential data.

Featured diodes are FD-100 (ultra fast switching computer diode) and FD-200 (high conductance, ultra fast "universal" type) with complete specifications and performance graphs. All diodes listed are immediately available in volume.

(Continued on page 180A)

Bendix Craftsmanship at work for you



NEW BENDIX® 6-OUNCE NOISE GENERATOR meets need for fast, accurate noise measurement in miniaturized package. It has special value on noise monitoring applications—such as microwave and radar receivers—where size, weight, and power drain rank equally important with band width. This Bendix model TN-1 is only 2.00″ x 4.25″ over-all, weighs a mere six ounces, features low power drain, and is ruggedly built for long, trouble-free service. Designed for transmission-type use over frequency range of 8500 to 9600 mc. For further information, write . . .

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NEWS New Products

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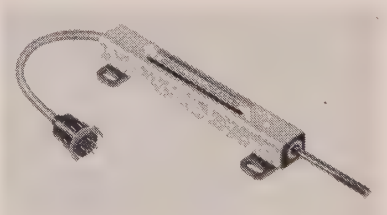
(Continued from page 178A)

Linear Motion Potentiometer

A new data sheet containing the latest specifications on their Models 155 and 156 Linear Motion Potentiometers is now available from **Bourns, Inc.**, 6135 Magnolia Ave., Riverside, Calif. Used primarily to detect the position of hydraulic actuators, pneumatic valves, positioning mechanisms, and linkage components, these components provide precision electrical signals proportional to mechanical position for recording, control and telemetering circuits.



The Model 156 Align-O-Pot® features a self-aligning shaft that permits measurement of linear displacement even when perfect alignment with the actuating member is impossible.

The Model 155 Liniometer® is a companion model, available for applications not requiring the self-aligning shaft feature.

Specifications for the Models 155 and 156 are: Travel: 0–1.5″ to 0–6.5″; Resistance: 1K to 50K; Resolution: as low as 0.001″; Linearity: ±0.5%; Operating temperature range: −55 to +175°C; Power rating: 1.0 watt per inch of coil at 40°C; Life expectancy; 1,000,000 cycles.

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NOT A MULTIMETER...
BUT...a new type of
INCREMENTAL
MULTIMETER!

● An entirely new type of test equipment, the Tensor #5880 Incremental Multimeter enables increments of any of the unit's ranges to be expanded about any point.

These ranges, including attachments, consist of *over 50 different* values of current, voltage, resistance, audio power, temperature and transistor characteristics with typical accuracies of 1.5% D.C. or 2% A.C. Increments as small as .01% can be measured and recorded!

The Tensor Incremental Multimeter may be employed wherever a strip chart recorder is used to display variations in a parameter that may be represented as a voltage, current or resistance capable of being measured on a Simpson "270" multimeter. Typical examples are the study of line voltage variations or power supply regulation.

Servo recorders of various sensitivities from 1 MV to 250 MV can be used. When used with a 5MV sensitivity external servo recorder expansions to ±1% of the full scale of the range are available. The attenuator enables the expansion to be adjusted to any value from 1% to 100%. A D.C. millivoltmeter may also be used if no permanent record is needed.

The Tensor Incremental Multimeter consists of a Simpson "270" multimeter, suppressor power supply, and attenuator. Package dimensions are 18" long, 4⅝" high and 7½" deep. It weighs 7 lbs.

PRICE $250 *DELIVERY FROM STOCK*

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a new name with a 50 year tradition in electronics

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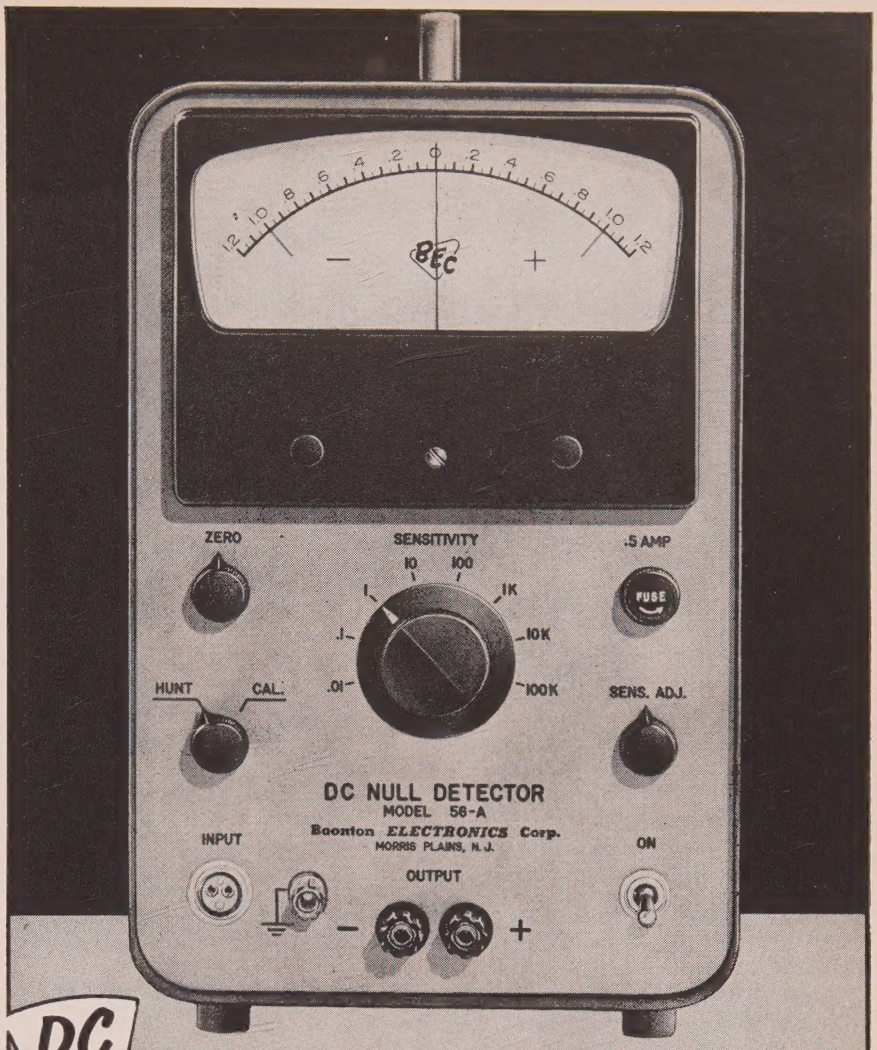

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HIGH INPUT IMPEDANCE 10 megohms constant on all ranges

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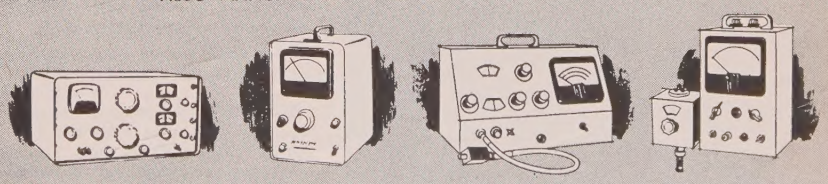
ALSO MANUFACTURERS OF THESE FINE INSTRUMENTS:



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0.1 Mc to 940.0 Mc

Determines resonant frequency of tuned circuits, antennas, transmission lines, by-pass condensers, chokes, etc. Measures inductance and capacitance. Also used as a signal generator, wave meter, frequency meter, and in many other applications.

This compact, lightweight grid-dip meter is available in the frequency ranges indicated.

Standard Model 59 complete	$168.00
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2.2 Mc - 420 Mc

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100 Kc - 4.5 Mc

Model 59-UHF
Oscillator
420 Mc - 940 Mc

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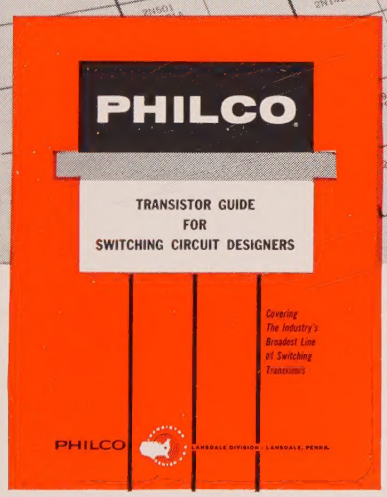

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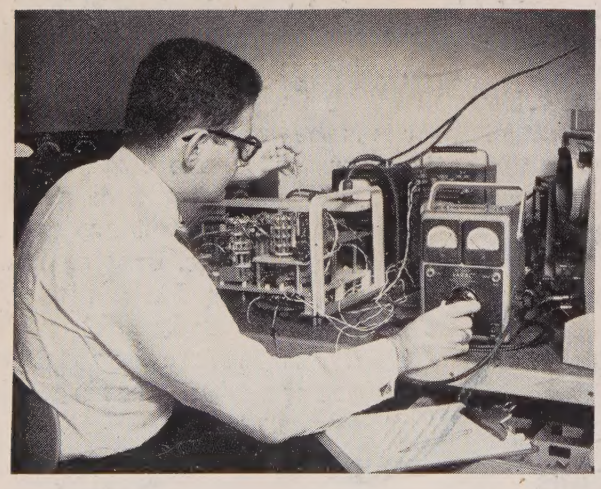

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